HONEYWELL

Ordnance Division

LOW INPUT VOLTAGE CONVERSION

REPORT NUMBER I

Contract Number DA-36-039-SC-90808

Department of the Army Project No. 3A99-09-001

First Quarterly Progress Report

1 July 1962 to 30 September 1962

U. S. Army Research and Development Laboratory,
Fort Monmouth, New Jersey

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Minneapolis-Honeywell Regulator Company
ORDNANCE DIVISION
Hopkins, Minnesota

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Contract Number DA-36-039-SC-90808

Prepared in accordance with Signal Corps Technical Requirement Number SCL-2101N, dated 14 July 1961.

Department of the Army Project No. 3A99-09-001

First Quarterly Progress Report

1 July 1962 to 30 September 1962

Object: **The object of this contract is to investigate all known methods of low input voltage conversion, determine the optimum approach, and build four converter models to meet specific requirements.**

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SECTION I

PURPOSE

The purpose of this contract is to investigate all known approaches and to determine the optimum methods of converting low voltage d-c power to higher voltage more usable d-c or a-c power for operating military electronic equipment. Specifically the most feasible methods of converting the low output voltages of thermoelectric, thermionic, solar cells, fuel cells or other source voltages in the range of 0.1 volt to 1.5 volts will be determined. All known or proposed methods including techniques and circuitry should be studied to determine the most feasible approach to obtain the optimum performance with respect to efficiency, regulation, life, ruggedness, weight, size, and ambient conditions. The compatibility of the conversion system with the power source output characteristics for proper impedance matching, regulation, and stability will be investigated. Specifically, effort will be directed towards achieving high efficiency with 75 percent efficiency as the tentative design goal.

The initial phase of this program is research and investigation to determine the optimum approach to the low input voltage conversion problem. The second phase is the fabrication of four converter models to prove that the application of the optimum approach is the most feasible solution to Specific Signal Corps Power Requirements.

SECTION II

ABSTRACT

During this first quarter, the Honeywell research and investigation team was organized and specific fields of investigation were assigned to engineering and research personnel having experience in these areas. A literature search was made to determine all known methods of power conversion and to obtain performance data on these methods and data on transducer devices. The following approaches were investigated in detail during this quarter:

1. Transistor Approach
2. Tunnel Diode Approach
3. Electromechanical Approach
4. Hall Effect Approach
5. Magnetoresistive Approach
6. Superconductive Approach
7. Photoresistive Approach.

Detailed investigation of the Liquid Metal Magnetohydrodynamic Approach was not made under this contract during this quarter because it appeared judicious to wait for results from a similar program being conducted by the Honeywell Research Center.

Calculations have been made to determine transducer requirements for each approach. Formulas have been derived and calculations made which determine the resistance ratios necessary between the "off" and "on" transducers in a push-pull circuit to achieve any given efficiency. This information has been used to determine requirements and feasibility of various approaches. A brief resume on the status of each approach follows:

TRANSISTOR APPROACH

Calculations have been made to determine the transistor parameters required to construct 50 watt push-pull converters operating from 1.5 volt and 1.0 volt sources with a conversion efficiency of 75%. These calculated parameters are beyond the state of the art of presently developed typical transistors. Some selected H-75 (Honeywell Type MHT-1902) power transistors might be obtained with the necessary parameters to achieve the desired performance in a two transistor converter; however, this may be difficult. Calculations have not been made for operation from sources of less than one volt because the severe requirements for operation at 1 volt and 1.5 volts would be quite difficult to overcome.

The development of special transistors may be necessary to obtain satisfactory converter performance. It may be possible to develop special transistors with the present state of the art if advantage is taken of the low voltage required for most parameters and if a low resistivity and thinner base material is used.

Methods of constructing converters with four or more transistors have been investigated. The use of four or more transistors in a 50 watt power amplifier appears to have more promise because the transistor requirements are reduced. The lower collector currents required for operation under this condition may allow the selection of transistors from presently developed devices. Thus it may be feasible to build a converter with a four-transistor power amplifier which can achieve 75% efficiency with a 50 watt output when operating from a 1.5 volt source. For a 150 watt output, 12 transistors would be necessary. The above statements are based upon calculations and experimental verification would be necessary for proof.

Transistor converter drive circuits were studied. These include self excited and separately excited circuits with voltage feedback and current feedback. Some form of current feedback appears desirable because of the wide range in input voltage and load fluctuations expected. Combinations of both voltage feedback and current feedback appear favorable.

A transistor converter having a separate oscillator to supply voltage drive to a four transistor power amplifier with current feedback appears to be the more optimum approach to obtain a 50 watt converter with desired performance characteristics.

TUNNEL DIODE APPROACH

Investigation of this approach shows that the main advantage of tunnel diode transducers is their low forward saturation resistance and their high forward current capacity. These devices must be switched between their optimum operating points to obtain maximum efficiency. The input voltage to a tunnel diode converter must be held within a narrow range in order to operate near the optimum points. Calculations have been made to determine tunnel diode parameters necessary to fabricate converters having overall efficiencies of 75% and 65%.

These calculations show that the tunnel diode parameter values required exceed the parameters that can be obtained from presently available devices. Thus tunnel diodes will have to be improved considerably before the desired efficiencies can be obtained. Improvement is necessary in the peak current to valley current ratio and in the valley voltage to voltage at peak current ratio. The voltage ratio requires the greatest improvement since this is low, causing high losses. It might be possible to improve this ratio with improved gallium arsenide semiconductor materials. Although this approach might be satisfactory if the devices are improved, it does not appear very favorable at this time. If a lower overall efficiency can be tolerated (about 50%) then this approach would appear more promising.

Further work will be done on the tunnel diode approach at a low rate of effort. This effort will be directed toward a more accurate determination of the state of the art and estimation of possible device improvements. It is known that several organizations are developing advanced high current tunnel diodes, and if the required high current and voltage ratios parameters are developed, effort on this approach will be increased.

ELECTROMECHANICAL APPROACH

Preliminary investigation of the electromechanical approach was directed toward determining the feasibility of a particular electromechanical configuration. This feasibility study showed that the particular configuration examined was gravity dependent and frequency limited. Thus the particular configuration did not appear favorable. Other electromechanical configurations may be feasible and further investigation will be directed toward determining other electromechanical configurations and their feasibility. The electromechanical approach is not a desired "solid-state" device but it will be investigated to determine its advantages, disadvantages, expected performance, and reliability.

HALL EFFECT APPROACH

Preliminary investigation has shown that the maximum attainable efficiency of a Hall effect transducer is only 1.7%. On this basis the approach appears unfavorable and has been ruled out.

MAGNETORESISTANCE EFFECT APPROACH

Calculations have been made on a magnetoresistive approach configuration. These initial calculations have not been conclusive because the approach appears marginal.

Our calculations indicate that achievement of the necessary "off" to "on" transducer ratio will be difficult at room temperature. Information indicates that the proper ratio might be achieved at liquid nitrogen temperatures, however. It will be necessary to study the magnetoresistive approach further to more accurately determine feasibility. The switching times and switching losses will also be considered in future calculations on this approach.

SUPERCONDUCTIVE APPROACH

Calculations have been made on several configurations and parameter requirements for the cryogenic approach. This approach obtains high "off" to "on" transducer resistance ratios by switching the transducer between its normal resistance state and its superconducting state at liquid helium temperatures. The machinery necessary to produce these low temperatures is complex. Calculations indicate that the transducer resistance ratios (including the lead resistance) should exceed 10, 000 to 1 in order to obtain the required efficiency. The fact that optimum lead dimensions must be used to minimize heat conduction through the leads places limitations on the design. Our calculations show that the major limitation of this approach is the lead heat conduction. Calculations show that the losses, including refrigeration losses, will be at least 5. 7 watts per ampere of current carried through the cryotron. This indicates that no net energy output can be obtained if the source voltage is less than 5. 7 volts. Actually the source voltage would have to be about 50 volts before any practical efficiencies could be obtained. Refrigeration could be furnished by a portable supply of liquid helium in place of obtaining refrigeration power from the device output. However, calculations show that the weight and volume of helium required for this possibility is excessive.

Thus our calculations have shown that this approach is not suitable for low input voltage conversion in the voltage range in which we are interested.

PHOTOCONDUCTIVE APPROACH

Investigation of the photoconductive approach has shown several limitations. Properties of various known photoconductive materials have been evaluated. The resistance ratios between the "off" (dark) transducer and the "on" (illuminated) transducer is a function of the operating frequency. The frequencies that will yield the desired ratios to achieve 75% efficiency are below 20 cycles per second. The method of supplying the required illumination and switching is a problem inherent in this approach. The photoconductive transducer necessary to handle our current requirements would require a large surface area and hence would be bulky. The primary limitation of this approach is the slow response. If more optimum photoconductive materials are developed, this approach might appear more promising than it does with present materials.

The results of these preliminary investigations have shown that the transistor approach appears feasible if a sufficient number of the proper transistors are used in the optimum configuration. The development of better transistors for this specific application is desirable. Further work will have to be done to evaluate the electromechanical and magnetoresistive approaches. The tunnel diode approach is marginal with present devices, but if the peak to valley ratios are improved, achievement of desired efficiencies with this approach may be feasible. Material improvement would be necessary to make the photoresistive approach feasible. The Hall effect and superconductive approaches are not feasible for our requirements. Effort will be directed at finding and evaluating other possible approaches to the low input voltage problem.

SECTION III

CONFERENCES AND LECTURES

On 27 June 1962, Mr. L. E. Alberts, Mr. G. Reiland, Mr. R. D. Fenity, Mr. W. K. Chaffee, and Mr. J. T. Lingle of Minneapolis-Honeywell visited Dr. E. Kittl, Mr. H. J. Byrnes, Mr. F. J. Wrublewski and Mr. W. L. Dudley at the U.S. Army Electronics Research and Development Laboratory, Power Sources Division at Fort Monmouth, New Jersey.

The Laboratory personnel outlined areas to be studied, and the proposed work for the program, particularly for the first quarter, was discussed. It was determined that the first 12 months of this program should be directed primarily toward determining the optimum approach. The remaining six months should be primarily devoted to model fabrication. It was indicated that more effort should be placed on approaches that are more promising or more unknown. Information should be obtained on each approach to determine the required transducer parameters necessary to achieve the desired efficiency. Aspects of future development should not be omitted but the main effort should be directed toward the most promising and feasible within the present state of the art. Several references to conversion system literature were obtained from the Laboratory for inclusion in the literature search.

Laboratory personnel suggested that progress reports should be arranged with each conversion approach contained in a separate section and each section signed by the personnel that performed the work.

On 11 July 1962 a conference was held at the Honeywell Research Center to formulate plans to investigate the various approaches. The following personnel attended this conference:

L. E. Alberts
R. D. Fenity
F. Exner
C. Motchenbacher
D. Long
O. Tufte
J. Garfunkel
J. Maupin
O. Lutes
J. T. Lingle

Each approach was discussed in detail and the program for the first quarter was outlined. The Honeywell Investigation team was organized and specific personnel were assigned to investigate the various approaches as shown in the organization chart on Figure 45.

On 25 September 1962 a conference was held at the Honeywell Research Center to review the first quarter's work and to discuss the report rough drafts for each approach. The personnel that attended this conference were the same as at the 11 July 1962 conference. The information discussed has been included in this report.

On the evening of 28 September 1962, Mr. R. D. Fenity and Mr. J. T. Lingle attended a Symposium of the American Ceramic Society in Minneapolis titled "The Role of Materials in Generation of Electrical Energy". Lectures were given on thermoelectric, thermionic, Fuel Cell, and Magnetohydrodynamic energy sources.

SECTION IV

FACTUAL DATA

This report is divided into separate sections which outline the investigation of each specific approach. Each section is signed by the personnel who performed the work. The work performed on the Literature Search including the Transistor Approach, the Tunnel Diode Approach, Electromechanical Approach, Basic Transducer Efficiency Calculations, Liquid Metal Magneto hydrodynamic Approach, Hall Effect and Magnetoresistive Approach, Superconductive Approach and the Photoconductive Approach is described in detail below.

A. LITERATURE SEARCH

To effectively initiate this program, a literature search has been made to determine all known methods of voltage conversion and to review work which has been previously done on the subject.

Information has been gathered on approaches, transducer materials, transducers, circuitry, efficiencies and power sources. Some of this information has been useful in determining the feasibility of various approaches. This particular information is referenced herein. A Bibliography of applicable literature has been compiled and is located in Appendix E. The literature which has been received and reviewed is indicated by the asterisk (*). Some of the articles have not yet been received but will be reviewed as they arrive.

B. TRANSISTOR APPROACH

The transistor approach to the low input voltage conversion problem falls into two areas; 1) transistor requirements and 2) transistor circuitry.

1. Transistor Requirements

The transistor parameters required to construct low input voltage converters have been calculated. Parameter calculations and assumptions for 1.5 volt and 1.0 volt inputs are shown in Appendix A.

a. Operation From a 1.5 Volt Source - The calculated requirements for operation from a 1.5 volt source are shown in Table I.

TABLE I

TRANSISTOR REQUIREMENTS FOR 1.5 VOLT OPERATION

Transistor Gain H_{Fe} = 40 Minimum
at I_c = 44.44 amps

Input Voltage V_{be} = 1.2 Volts Maximum
at I_b = 1.33 amps
I_c = 44.44 amps

Forward Saturation Voltage V_{ce} (Sat.) = 0.133 Volt Maximum
at I_c = 44.44 amps
I_b = 1.20 amps

Pulse Switching "on" time*= 20 μ sec. Maximum
at I_c = 44.44 amps
I_b = 1.8 amps

Pulse Switching "off" time* = 20 μ sec. Maximum
at initial I_c = 44.44 amps
V_{be} = 1.2 volts
I_b = .5 amp

TABLE I (Cont.)

Voltage Ratings Required. Collector-to-base Voltage V_{CB} = -8 volts Collector-to-emitter Voltage V_{CE} = -8 volts (both forward and reverse emitter bias) Emitter-to-base Voltage V_{EB} = 3 volts Collector Current I_C = -65 amps dc Minimum Base Current I_B = -5 amps Thermal Resistance Junction to mounting base Θ_{J-MB} = .5° c/watt Maximum
*Rise and fall times are interpreted here in terms of switching time under the stated bias and back bias conditions. The "switching off time" shall also include the storage time if the device comes out of saturation during the storage interval.

The above specifications show the calculated requirements necessary to fabricate a push-pull converter utilizing a single pair of transistors and to achieve 75% efficiency. To achieve this, all requirements specified must be met. If it is difficult to meet one requirement, desired results might possibly be obtained by making considerable improvement in other requirements to balance out marginal performance on the one requirement. It must be pointed out, however, that the trade off possibilities are very limited, especially where V_{ce} (saturation), drive power required, and switching times are concerned.

These calculations are based upon rather optimistic assumptions for other circuit losses; therefore, the transistors should at least meet the above requirements. We desire to boost these requirements if possible. Specially selected H-75 type transistors might possibly meet the above requirements; however, fabrication of special transistors may be necessary to provide the desired device. This might be accomplished by using a lower resistivity germanium and by making the base layer thinner. It can be noted that the V_{CB}, V_{CE}, and V_{BE} required voltage ratings are very low and it is anticipated that advantage can be taken of this fact. The above calculations show that the transistor losses consist of the following:

Switching losses	=	2.59 watts
Drive losses	=	3.20 watts
Quiescent I_C V_{CE}		
Saturation loss	=	5.90 watts
Total transistor loss		11.69 watts

If the estimated losses are expressed as a percentage of converter input they are:

$$\text{Switching loss} = \frac{2.59}{66.7} = 3.89\%$$

$$\text{Drive loss} = \frac{3.2}{66.7} = 4.80\%$$

$$\text{Saturation loss} = \frac{5.90}{66.7} = 8.85\%$$

Total transistor loss = 17.54 % of converter input

Notice that the quiescent I_C V_{CE} saturation loss is the greatest loss and is about twice the drive loss. Therefore, to obtain performance improvement, the most effort should be concentrated on this loss. The next highest loss predicted is the 3.20 watt drive loss. This consists of 1.6 watts actual drive plus an additional 1.6 watts consumed in the drive circuit to achieve an optimum type of drive wave form. This loss may be consumed in drive limiting resistors and/or a separate saturating feedback transformer. It might be possible to reduce this assumed 1.6 watt loss by various circuit arrangements; however, this may result in additional switching losses tending to cancel any reduction in drive losses. Because of this, the drive losses are probably optimistic. The switching losses are based upon the assumption that optimum drive provided will over drive the transistor during the switching interval and achieve rapid switching to minimize switching losses. The assumption of 20 μ sec. switching time is believed to be realistic for this size transistor operating at low input voltages and high currents. The switching characteristics assumed are also considered realistic. The achievement of an optimum ratio between the switching losses and the drive losses is desired. For the purpose of calculating transistor parameter requirements, the above switching and drive losses are considered optimistic. These losses are based upon a minimum gain requirement of 40 at 44.4 amperes which may also be optimistic.

b. <u>Summary of Requirements for Operation From a 1.5 Volt Source</u> - The above calculations outline the transistor parameters necessary to achieve 75% efficiency with a two transistor push-pull converter operating from a 1.5 volt source. Examination of these parameters shows that presently available transistors and specifications do not meet these requirements. Some specially selected Honeywell type H-75 transistors might approach these requirements. It appears very likely that special transistors could be fabricated to meet these requirements with the present state of the art provided advantage is taken of the very low voltage requirements for V_{CB}, V_{CE}, and V_{EB}. The most important parameter is deemed to be V_{CE} (sat.) because this contributes the greatest loss.

c. Operation From A 1.0 Volt Source - The calculated requirements for operation from a 1.0 volt source are shown in Table II.

TABLE II

TRANSISTOR REQUIREMENTS FOR OPERATION FROM A 1.0 VOLT SOURCE

Transistor Gain H_{FE} = 60 minimum
at I_C = 66.7 amps

Input Drive Voltage V_{be} = 1.2 volts maximum
at I_b = 1.33 amps.
I_c = 66.7 amps.

Forward Saturation Voltage V_{CE} Sat. = .078 volt maximum
at I_c = 66.7 amps.
I_b = 1.20 amps.

Pulse Switching "On" Time * = 20 μ sec. maximum
at I_c = 66.7 amps.
I_b = 1.8 amps.

Pulse Switching "Off" Time * = 20 μ sec. maximum
at initial I_c = 66.7 amps
I_b = .5 amp
V_{be} = 1.2 volts

Voltage Ratings Required:

Collector-to-base Voltage V_{CB} = -8 volts
Collector-to-emitter Voltage V_{CE} = -8 volts
(both forward and reverse emitter bias)
Emitter-to-base Voltage V_{EB} = 3 volts
Collector Current I_C = -66.7 amps minimum
Base Current I_B = -5 amps
Thermal Resistance
Junction to Mounting Base Θ_{J-MB} = .5° c/watt maximum

* Same notation for rise and fall times as indicated in Table I.

The above calculations show the required transistor parameters for operation from a 1.0 volt source. These requirements are much more severe than the requirements for the 1.5 volt source because required transistor collector currents and gains are higher. Further, the saturation voltage must be much lower at these greater collector currents. Examination of recent transistor specifications shows that no presently available transistors approach these requirements. The feasibility of fabricating transistors to meet these requirements will be investigated.

Satisfactory converter operation from source voltages below 1.0 volt will be more difficult to achieve than from a 1.0 volt case. Calculations have not been made for transistor converter operation below 1.0 volt because the 1.0 volt case does not appear sufficiently promising.

Two selected Honeywell Type MHT1902 (H-75) transistors have been obtained and are being tested to verify that selected transistors might satisfy requirements for converter operation from a 1.5 volt source.

2. Transistor Circuitry

a. Increased Power Capability and Improved Performance by Incorporating More Power Transistors - Above calculations show that a two transistor converter operating to required performance specifications from lower voltage sources requires transistor characteristics presently beyond the state of the art. Because of this, the feasibility of using some parallel combinations of transistors with less stringent requirements should be considered.

One of the better methods consists of operating the effective inputs to two converters in parallel and connecting the output secondary windings from each converter in series or in parallel. This would reduce the necessary transistor collector current to one half that required for transistors used in a single, push-pull converter. Using this scheme, drive power is furnished to four transistors in place of two. If the transistor gain can be doubled at these lower current levels, then the required drive power will remain approximately the same. Losses due to back bias and leakage in the "off" position would remain approximately the same for each transistor, and hence a device using four transistors when compared to a two transistor device would have this loss doubled. This loss is very small compared to other losses. The advantage of using more transistors is that the required V_{CE}(Sat) can occur at lower collector currents. Further, a possible increase in gain at lower collector currents is likely so that some overall drive reduction may be expected. A comparison can be made for the 1.0 volt input voltage case. If four transistors are used, the required collector current for each transistor is $\frac{66.7}{2}$ or 33.35 amperes. The required V_{CE}(Sat.) is still .078 volt maximum. This of course will still be difficult to achieve even at 33.35 amperes but is easier at this level than the higher current level. The assumed gain for the 1.0 volt case was 60. This would have been difficult to attain at 66.7 amps I_C but may be reasonable at 33.35 amps I_C. This comparison shows that the use of four transistors reduces transistor requirements considerably, but with a one volt source, the requirements still exceed published transistor specifications. It is possible that selected units can be obtained to give satisfactory performance or additional transistor pairs might be used. In addition, special transistors might be fabricated for this purpose.

Consider a four transistor converter operating from a 1.5 volt source. The required collector current per transistor is $\frac{44.44}{2}$ = 22.22 amperes. The required V_{CE} (Sat.) for this case is 0.133 volt maximum. The preliminary specifications for the Honeywell H-75 type transistor indicate a maximum V_{CE} (Sat.) of 0.25 volt at 25 amperes I_C. Thus it would appear to be entirely possible that specially selected units with a V_{CE} (Sat.) of .133 volt at 22.22 amps might be obtained. Further, the gain for this transistor was assumed to be 40. The tentative specifications for the H-75 list the minimum gain of 40 at

25 amps. Thus we could expect to obtain selected units having both the required low V_{CE} Sat. and a gain considerably higher than 40, perhaps in the 60 to 80 range at 22.22 amperes. Because of this, fabrication of a 50 watt converter operating at 75% efficiency from a 1.5 volt source appears feasible if four transistors are used. If a 150 watt device were desired it might be feasible to do this with 12 transistors.

b. Construction of Converters with More Than Two Transistors - The methods of connecting four transistors into a converter circuit to divide the maximum collector current between two or more transistors must be considered. In order to divide the collector current, the effective primary circuits should be connected in parallel with the source as shown in Figure 1.

These converters could conceivably be operated from two isolated power sources as shown in Figure 2.

The possibility of two separate power source modules might be considered for redundancy to increase reliability of the power source. If the Figure 2 arrangement were used, a short circuit failure in one of the converter primary circuits would not short out the second power source and hence the device should be capable of delivering half power. Thus Figure 2 provides redundancy for both power source failure and converter primary section failure. A third method of connecting additional transistors into the circuit might be simple parallel operation of transistors as shown in Figure 3.

In this circuit, resistors R_1, R_2, R_3, and R_4 have been added in series with the emitter circuits to prevent parasitic oscillations. This will tend to equalize the load among the parallel transistors. These resistors will introduce additional losses in the circuit; therefore this arrangement is deemed unsatisfactory for maximum efficiency, high current low voltage operation. Thus, the primary circuit arrangements of Figure 1 and Figure 2 appear to be the more promising.

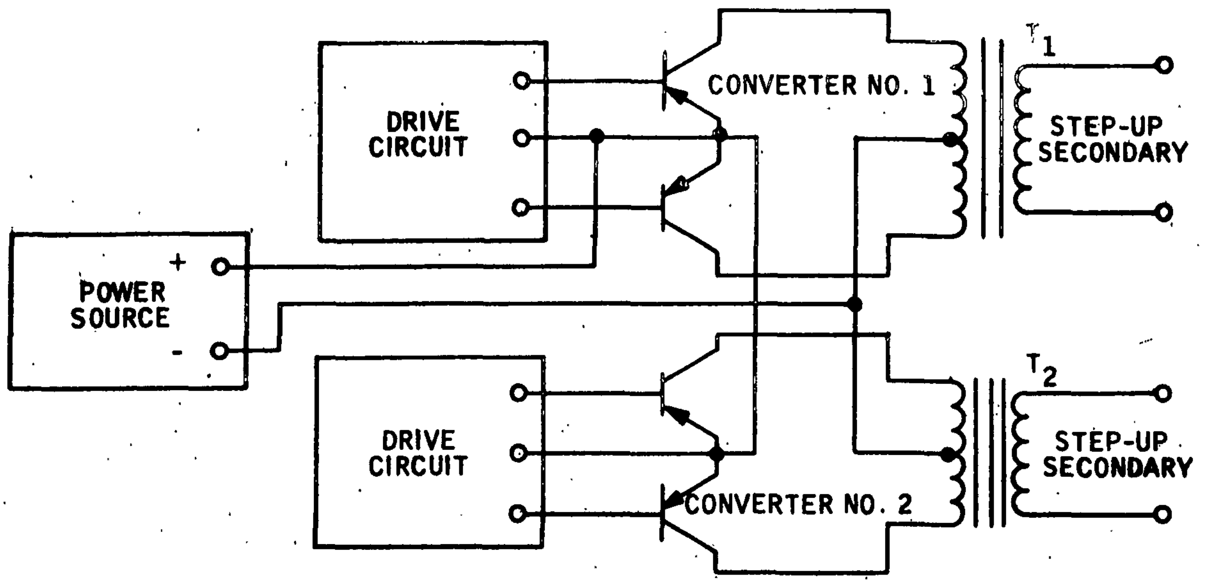

Figure 1 - PARALLEL PRIMARY CIRCUITS

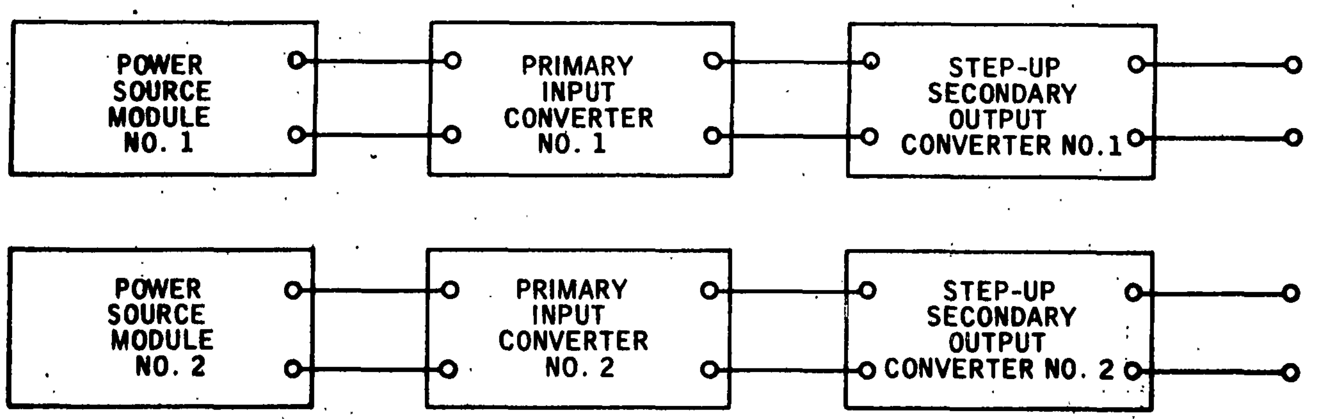

Figure 2 - ISOLATED SOURCES

There are two methods of connecting the secondary circuits, series and parallel. In the series connection each secondary may step up the voltage to half the desired amount as shown in Figure 4.

In the Figure 4 arrangement, identical secondary current I_o will flow through each secondary winding tending to maintain equal load currents in each converter. With equal secondary currents, the transistor collector current in Converter No. 1 will be maintained nearly equal to the transistor collector current in Converter No. 2. The primary advantage of this primary and secondary circuit arrangement is maintenance of equal load currents in each converter. The necessity of equal load currents, however, is questioned because it is anticipated that the relatively high source impedances expected in the power source may limit the collector currents. Since the voltage is very low, small differences in V_{CE} (Sat.) may be more important than equalization of collector currents.

The effect of this arrangement on the switching characteristics and on switching losses must be examined. The slowest switching transistor might control the switching time. With this arrangement, the step up ratio of the transformer is reduced which affects transformer design and the values of reflected impedance.

The second method of connecting the converter secondaries is shown in Figure 5.

In the Figure 5 arrangement, the rectified secondary outputs are connected in parallel across the load. Each converter delivers its own power to the load independently except for coupling through the common power source and common load. Thus with this connection, one converter can still furnish load current if the other converter fails--providing that the failure does not short either the load or the source. One of these converters could supply the normal output voltage at half load. With parallel secondaries, the switching in each is

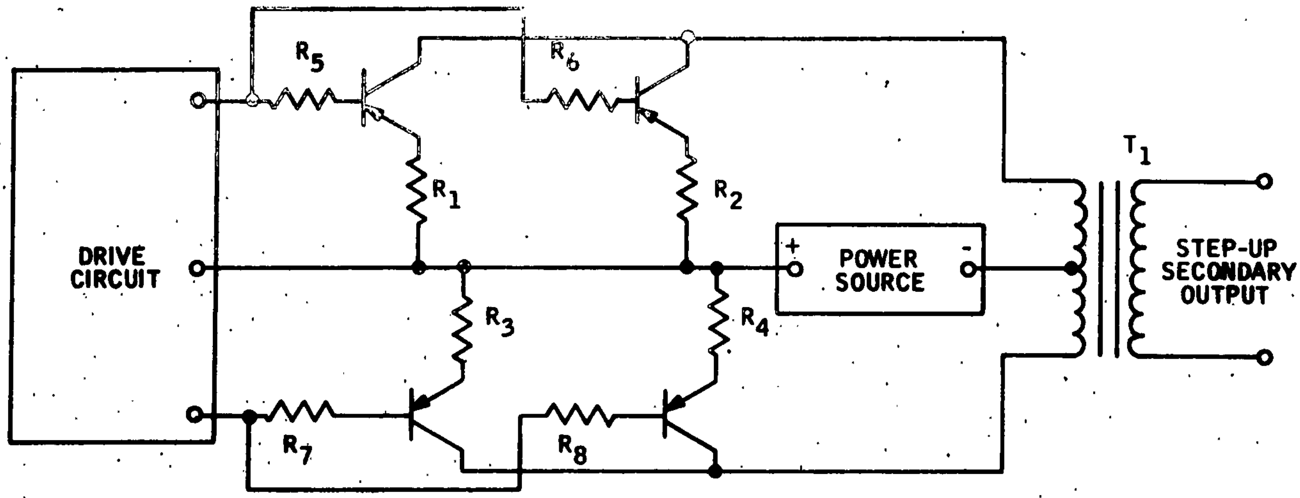

Figure 3 - PARALLEL TRANSISTORS

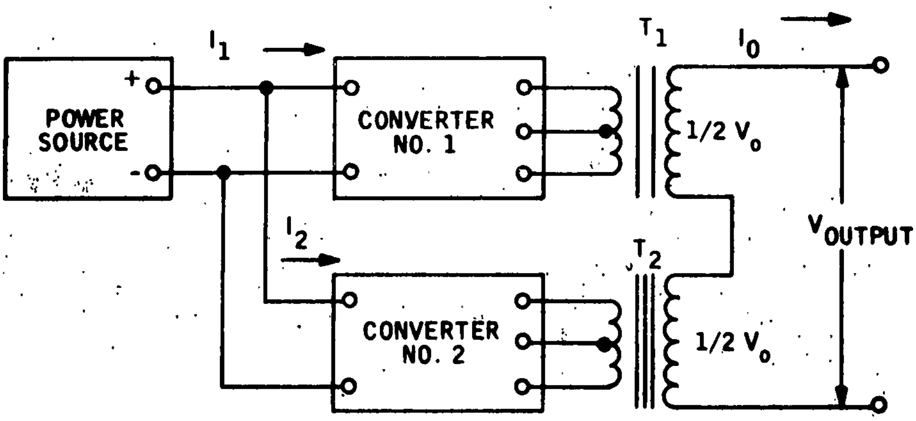

Figure 4 - SERIES CONNECTED SECONDARIES

independent of the other and thus slower switching time of one converter should not slow the other. One of the important features of parallel secondary circuits is that the two converters can be made to switch at different instants. If the converters are phased 90° apart (or some other angle) then one converter will be fully "on" and overdriven while the second converter is switching. This allows the saturated transistor to immediately pick up a considerable portion, if not all, of the load formerly carried by the switching transistor. The transistor storage effect and the fact that the saturated transistor is overdriven should allow the conducting transistor to pick up the entire load. This would allow the switching transistor to switch at extremely light load currents. Utilization of such a scheme should diminish switching losses because switching current and switching time would be reduced considerably. This approach appears attractive in the low voltage--high current converter because 75% efficiency is difficult to obtain. Reduction in switching losses may allow operation at higher frequencies to achieve weight reduction. Ripple frequency is also doubled reducing filter requirements. This arrangement does not insure equal division of the load current between the two converters; however, with high source impedance, equal division of the loads may not be particularly important because the transistors are operating considerably below their power ratings. With a low source voltage the most important factor is load division which minimizes voltage drops. Parallel connection of the secondary circuits appears to have more advantages than series connection in this application because the following are provided:

1. Greater reliability through redundancy,
2. Possible reduction in switching losses,
3. Possible increase in ripple frequency would reduce filter weight,
4. Possible higher optimum operating frequencies may reduce weight of transformers and filter.

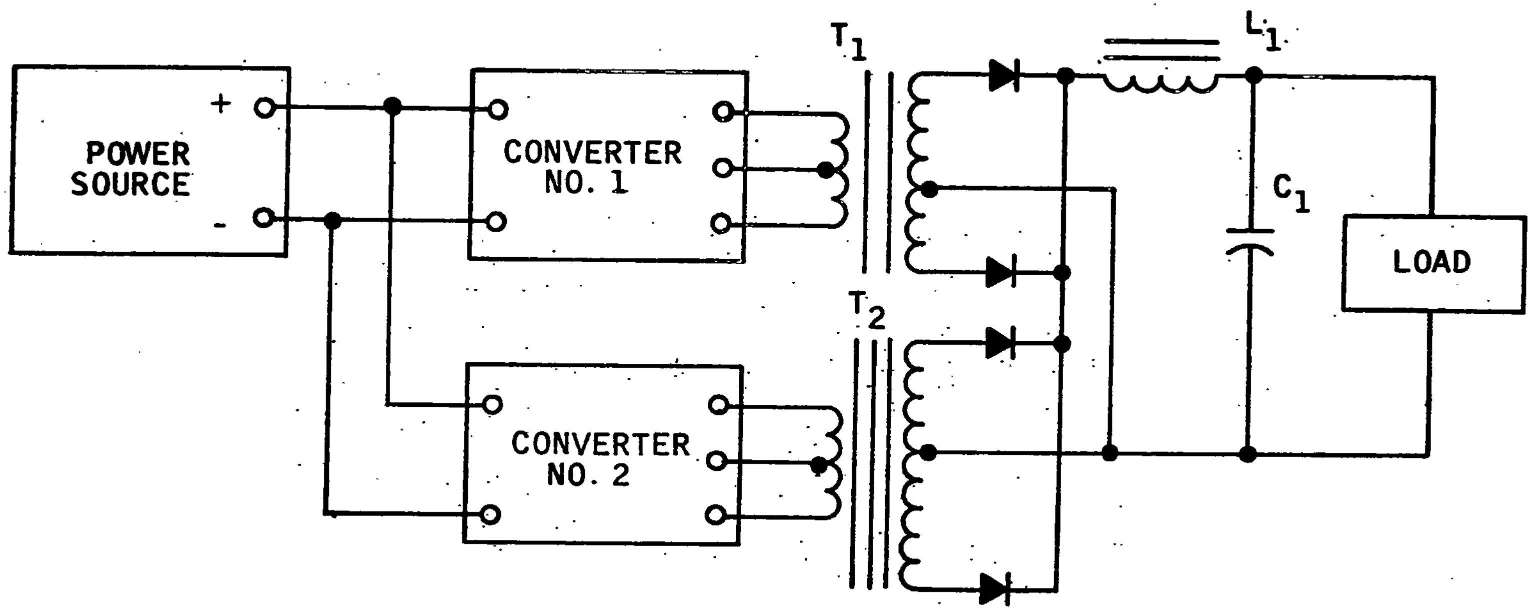

Figure 5 - PARALLEL CONNECTED SECONDARIES

Some additional circuitry might be required to synchronize two converters out of phase. The advantages gained by this type of operation may offset increased complexity. In order to determine the switching losses of a device utilizing two converters operating 90 degrees out of phase, breadboard construction would be desirable to obtain switching loss and switching speed data. This breadboard should be operated from a single source as well as two separate sources. Resistors would be placed in series with the source to simulate high source impedance and to determine the effects of high source impedance on the switching characteristics.

c. Drive Requirements for Transistor Oscillators - Several types of feedback drive circuits may be used for transistor oscillators. Two main categories are voltage feedback and current feedback.

1) Voltage Feedback - Voltage feedback circuits derive their power from a winding on the output transformer either directly or through an impedance and a second feedback transformer. Examples of this type of drive are the Uchrin-Royer circuit shown in Figure 6 and the Jensen circuit shown in Figure 7. In both circuits the feedback voltage signal is proportional to the input voltage. Thus with a low input voltage the available drive voltage and current is low but the output voltage is also low.

Consequently the output current and collector current are also low if the load impedance is fixed. These oscillators tend to continue operation over wide input voltage ranges because the drive is adequate with fixed load impedance. Due to the non-linear transistor input characteristics these oscillators do experience some difficulty at extremely low voltages because of a rise in input impedance. Sufficient drive should be available to provide operation at the lower design voltage limit and over the ambient temperature range. Fulfilling this requirement results in a degree of over drive and higher drive losses. Over drive tends to reduce the transistor

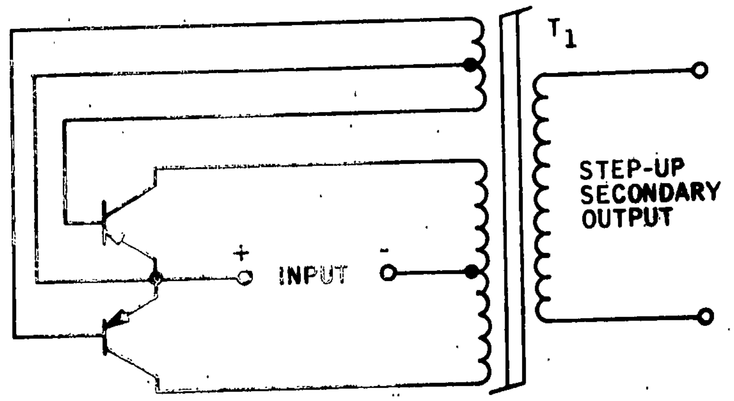

Figure 6 - VOLTAGE FEEDBACK (UCHRIN-ROYER CIRCUIT)

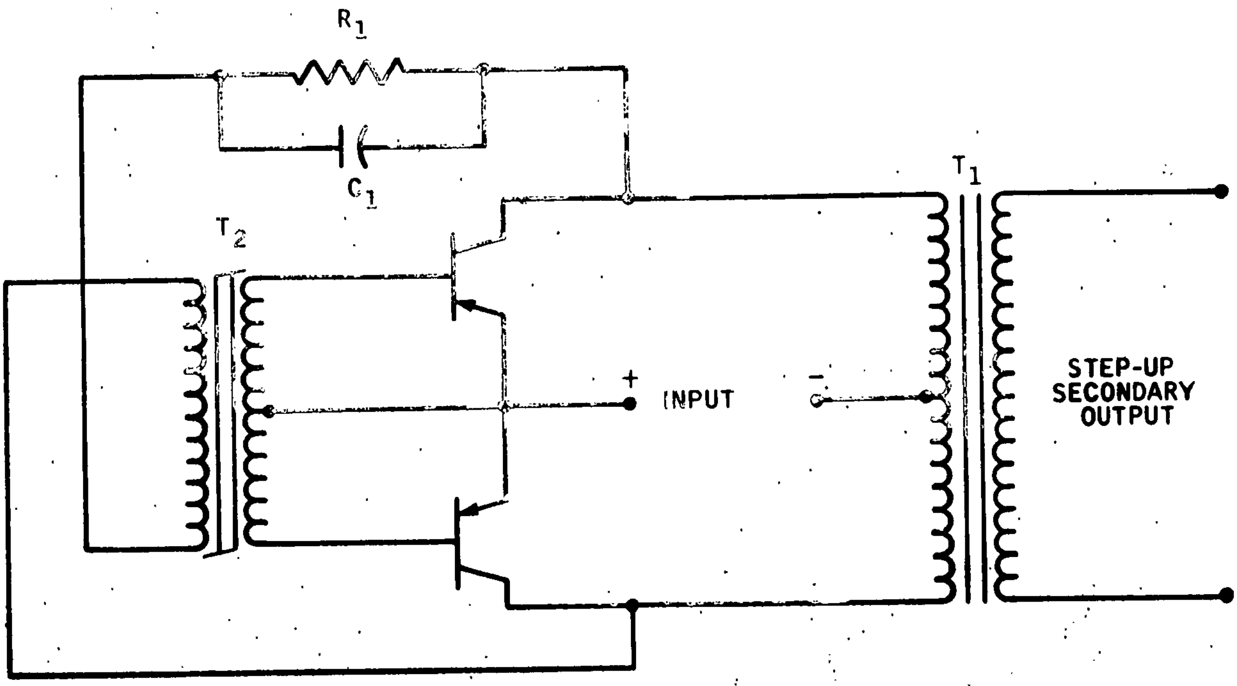

Figure 7 - VOLTAGE FEEDBACK (JENSEN CIRCUIT)

saturation voltage loss somewhat. For optimum performance the drive must be optimum. Resistors can be used in series with the transistor bases to limit the base drive currents at higher input voltages, but these also contribute to circuit dissipation.

The circuit of Figure 6 is the simpler because it has few components. This circuit accomplishes switching by saturation of the power transformer at the end of each half cycle. Core saturation increases the transistor collector current to the point where drive current is insufficient causing the transistor to come out of saturation. This causes the transistor voltage drop to increase reducing the transformer and feedback induced voltage which initiates the switching action. This circuit has the following disadvantages:

1. Switching occurs at high collector currents causing high transient switching losses.
2. Necessity of adequate drive for all conditions results in high collector currents and high switching losses even in the light load condition.
3. The output transformer must be saturated and hence torroidal square loop cores are desirable to reduce core losses.
4. Because the output transformer is saturated the effects of "apparent conversion of core loss current and load current into magnetizing current"* can become appreciable if the secondary current declines more rapidly than the primary current. This effect leads to high switching losses and must be minimized by optimization of the secondary leakage inductance parameter.
5. Saturation of the output transformer fixes many of the design parameters making this type of device more difficult to design and less flexible.

* This effect is defined in "Power Transistor Circuitry" Quarterly Progress Report III, pp. 21-23, by Minneapolis-Honeywell Ordnance Division, February 15, 1957.

This circuit has simplicity as its prime advantage and is widely used because of this fact. In general these devices are short circuit protected because a completely shorted secondary removes the feedback stopping the oscillator.

The circuit of Figure 7 utilizes a small saturating transformer and an impedance connected in series across a winding on the output transformer. In this circuit the smaller feedback transformer saturates at the end of each half cycle. Feedback transformer core saturation increases feedback current sharply causing most of the voltage across the feedback network to be dropped across the series impedance. This reduces the feedback drive voltage sharply causing the transistor drive to be switched. One of the important features of this circuit is that the transistor collector currents do not increase appreciably during the switching interval because the series impedance limits current through the feedback network during saturation. Thus the transistors are switched by drive control at normal collector currents rather than by loading beyond available drive. Because the transistors are switched at lower collector currents, transistor switching losses are lower with the Figure 7 circuit than with the Figure 6 circuit.

The Figure 7 circuit has the following advantages:

1. Switching losses are low because transistors are switched by drive control at normal collector currents and switching speed is high due to rapid voltage drop across series impedance.

2. Less difficulty is encountered with "apparent conversion of core loss current and load current into magnetizing current"* during switching interval; lower primary currents during switching, and the series impedance reduce this effect. Since the output transformer is not saturated, energy stored in the primary leakage inductance is less.

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* See note on page 26.

3. Transformer losses can be lower because a small transformer is saturated in place of the output transformer.

4. Saturation of a small feedback transformer and the use of a series impedance allow greater flexibility in frequency selection and in synchronization of operating frequency with external signals.

5. Circuit tends to be self-protecting against short circuits since short removes drive.

6. Transformer construction can be conventional because the output transformer is not saturated. This provides greater flexibility for optimizing leakage inductance and other parameters.

7. This circuit is more flexible and easier to design because the feedback and frequency control circuit is separate from the output transformer.

8. Over drive during switching can be provided by placing a small capacitor across the series impedance. This allows considerable increase in transient switching drive producing more rapid switching with lower losses.

This circuit has as its prime disadvantages the increased complexity of an additional transformer and the series impedance.

2) <u>Current Feedback</u> - Current feedback drive is obtained from a current transformer placed in series with the load current path. This small transformer can be placed either in the primary circuit or in the

secondary circuit. It is usually made to saturate in order to initiate the switching action. Figure 8 shows a current feedback converter with the current transformer in the primary circuit. Figure 9 illustrates a converter with the current feedback transformer in the secondary circuit. The prime advantage of the current feedback circuits is that drive current is proportional to load current and hence drive losses are reduced when operating at light loads.

One notable application for this type of circuit is the charging of photoflash capacitors from portable batteries in a period of a few seconds. In this application, high initial charging currents can be supplied because the proportional current feedback will insure operation of the transistors in the low impedance saturation region at high collector currents. As the capacitor becomes charged, the load becomes less and drive power is reduced to provide operation at higher efficiency for low current drain during standby. This type of operation improves battery life. Thus it can be seen that current feedback is desirable to provide optimum drive and high efficiency for devices which have wide variations in load current.

The Figure 8 current feedback circuit introduces additional voltage drops in the primary. The transistor input impedance and other impedances in the drive circuit are reflected into the primary circuit by the square of the turns ratio. The current transformer turns ratio required, $\frac{N_{1P}}{N_{1S}}$, equals the reciprocal of the assumed operating gain, $1/h_{FE}$. By assuming values of transistor gain and input impedance, this reflected voltage drop magnitude can be estimated. Assume the following conditions:

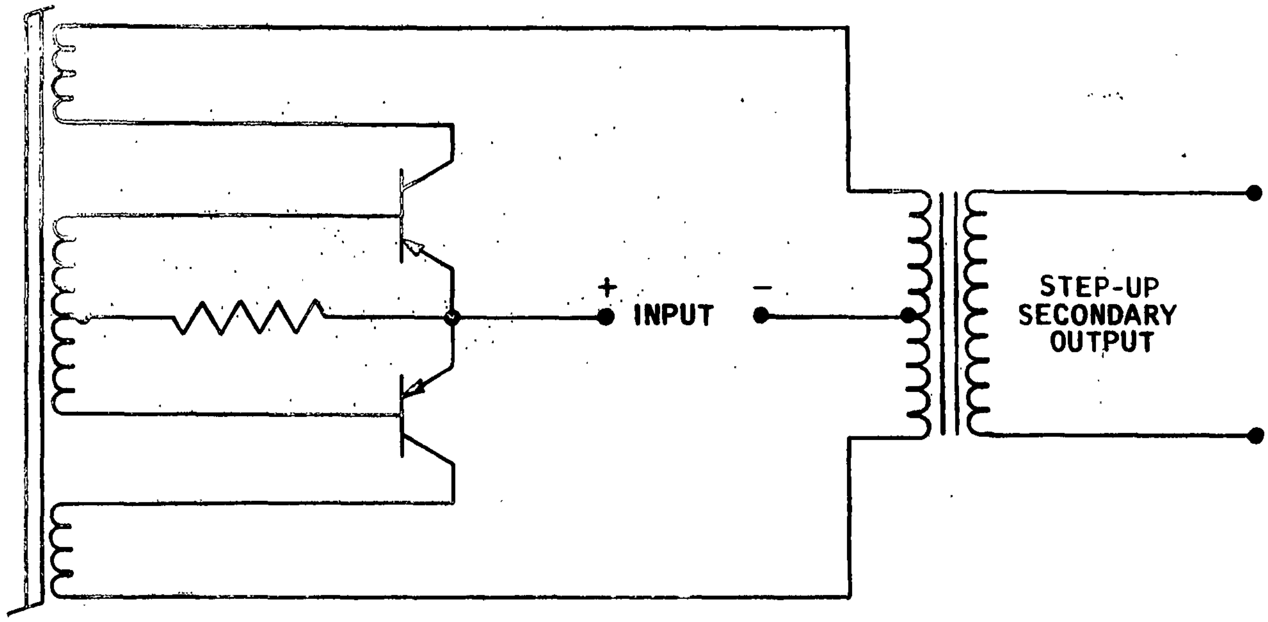


Figure 8 - HONEYWELL CURRENT FEEDBACK CONVERTER

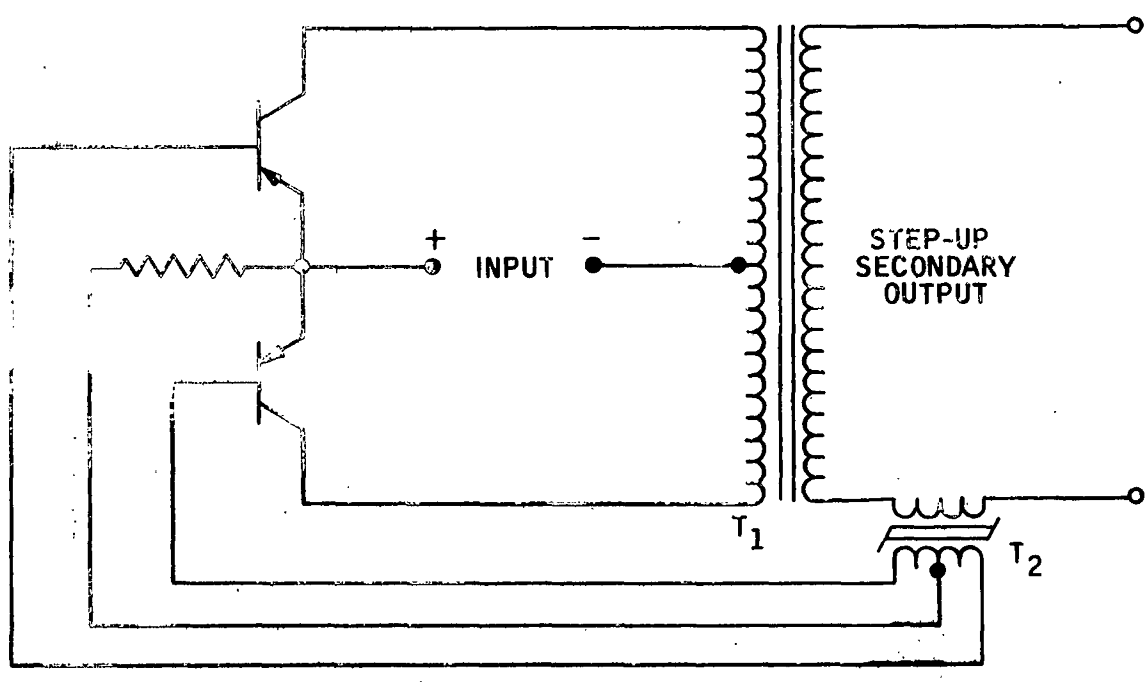


Figure 9 - PERLMAN CURRENT FEEDBACK CONVERTER

1. Transistor base current = 1 amp
2. Transistor collector current = 20 amps
3. V_{eb} = 1 volt.
4. R'_b = Reflected impedance.
5. $R'_b = R_b \left(\frac{N_{1P}}{N_{1S}}\right)^2$; but $\frac{N_{1P}}{N_{1S}} = 1/h_{FE}$; therefore,
6. $R'_b = R_b / (h_{FE})^2$

Thus the transistor input impedance, R_b, equals 1 volt/1 amp or 1 ohm. The current feedback transformer turns ratio is $1/h_{FE}$ or I_b/I_c = 1 amp/20 amps = 1/20. The impedance reflected into the primary circuit then becomes $1/20^2 \times R_b$ or $R_b/400$ = 1/400 ohm. With a collector current of 20 amperes the voltage drop across this reflected impedance would be I_c (1/400 ohm) or 20/400 which equals .05 volt. This .05 volt drop at 20 amps then furnishes the required 1 watt drive power. It has been previously estimated that the maximum allowable transistor saturation voltage drop V_{CE} is .133 volt at 22.22 amperes. The drive power is $\frac{20 \text{ amps} \times .05 \text{ volt}}{20 \text{ amps} \times .133 \text{ volt}}$ or a magnitude of 37.6 % when compared to the transistor saturation loss. If the source voltage were higher, the net effect of this reflected impedance would be less, but with low source voltages impedances in the primary circuit must be minimized. This estimate did not include resistance of the current feedback transformer winding which also should be considered. It is anticipated that this winding will consist of one turn. If additional current drive is needed, it may be necessary to change the transformation ratio which might increase the drop across the primary winding.

The circuit of Figure 9 incorporates the current feedback transformer primary winding in the output transformer secondary circuit. Since the secondary current is less than the primary current, the current transformer turns ratio will be less. The current transformer turns ratio will be determined by the desired transistor gain and the primary to secondary turns ratio of the output transformer. Thus $\frac{N_{2S}}{N_{2P}} = \frac{h_{FE}}{N_{1S}} \cdot N_{1P}$. The equivalent transistor input impedance reflected into the secondary is then $R_b \left[\frac{N_{1S}}{h_{FE} N_{1P}} \right]^2$.

In order to compare this reflected impedance directly with the circuit of Figure 8, the equivalent secondary impedance should be multiplied by the square of the primary to secondary turns ratio. Both circuits are then referenced with respect to the primary. This then becomes

$$R_b \left[\frac{N_{1S}}{h_{FE} N_{1P}} \right]^2 \cdot \left[\frac{N_{1P}}{N_{1S}} \right]^2 \quad \text{or} \quad \frac{R_b}{(h_{FE})^2} .$$

Thus it can be seen that the transistor input impedance reflected into the primary in series with the transistor will be the same regardless of a primary or secondary current transformer location. The only difference affecting the location of the current transformer is the power transformer efficiency and leakage inductance parameters. Since the efficiency is high, this will have little effect. The leakage inductance parameters, however, may affect the converter switching time.

Another factor affecting the choice of current feedback transformer position is the type of filter used in the output. If a single current feedback converter having the current transformer in the secondary circuit were used with a choke input filter, as shown in Figure 10, the choke would tend to maintain the flow of secondary current during the switching interval. This continued secondary current would keep the current feedback transformer energized and would tend to cause slow switching.

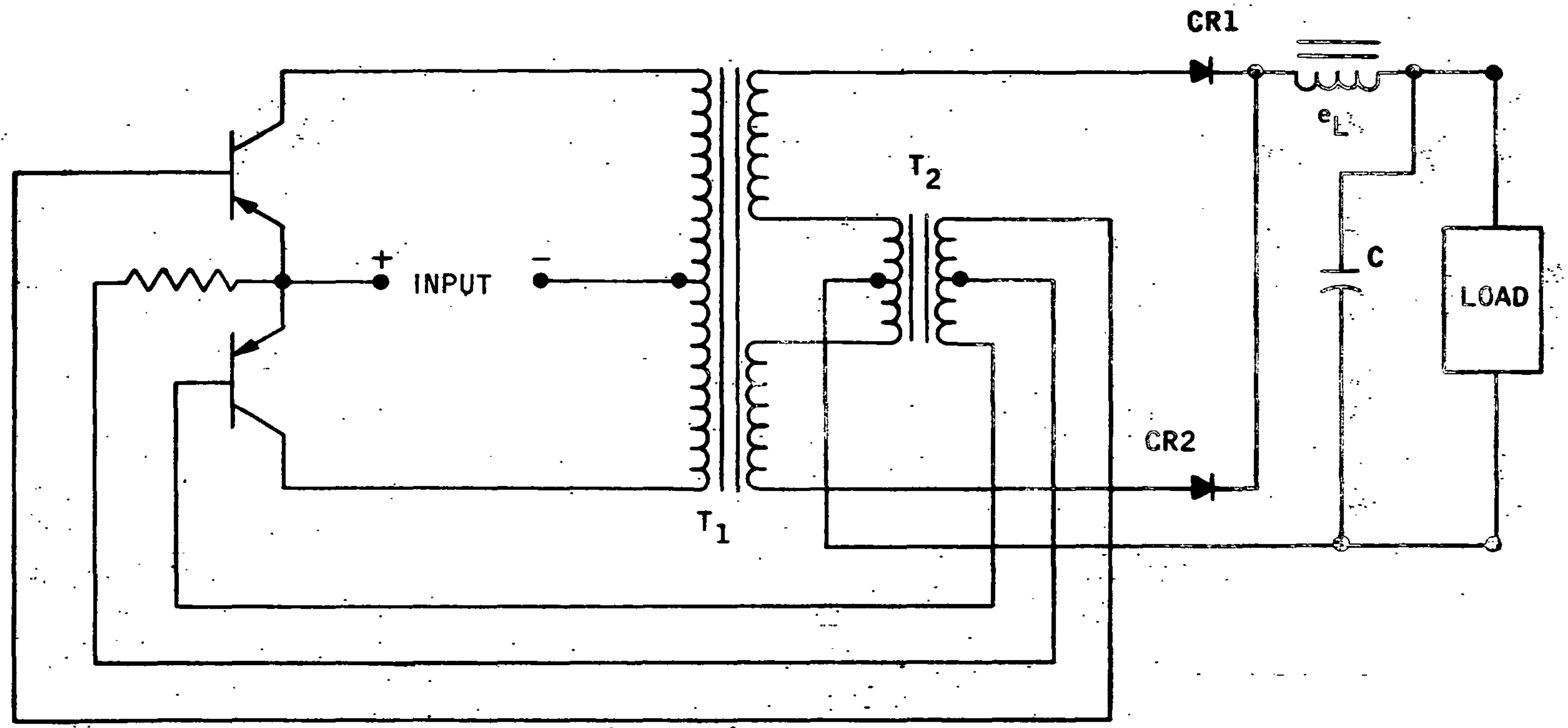


Figure 10 - CURRENT FEEDBACK CONVERTER WITH CHOKE INPUT FILTER

Since this is undesirable, use of a secondary current transformer with a choke input filter does not appear desirable. On the other hand, if two converters were operated out of phase with parallel secondaries into a common choke input load, the use of current transformers in the secondaries may be satisfactory. In this case, secondary and feedback current would not be maintained during the switching interval because the second converter would pick up the load. The use of a current transformer in the secondary circuit with a capacitor input filter provides satisfactory switching because the secondary current declines rapidly to zero when the secondary voltage drops below the capacitor voltage. Converters with current feedback transformers in the primary power circuit also show satisfactory performance with capacitor input filters.

d. Converter Equivalent Circuits - The equivalent circuit of a self-excited converter with a voltage feedback is shown in Figure 11. It can be noted that current to supply the voltage drive must flow through the transistor and the primary winding. The transistor must have sufficient capacity to carry the necessary additional current for drive as well as the current to supply the load and losses. This additional drive current, although small, does tend to shift the transistor characteristics to slightly lower gain and causes slightly higher saturation voltage drops. A small portion of the primary volt-amperes are used to supply the drive and the remaining portion is transformed to the secondary circuit. It is desirable to transform the drive power to the transistors with high efficiency so as to minimize drive requirements, additional transistor saturation losses, and necessary capacity.

Figure 12 shows the equivalent circuit for a self-excited current feedback converter. As in the above case, the current required to supply the drive, the load, and the losses must flow through the transistor. All of this current also flows through the reflected impedance of the current transformer R_{FB}, X_{FB}. The power consumed in the voltage drop across this impedance is transformed by the current transformer to supply the transistor drive. This of course

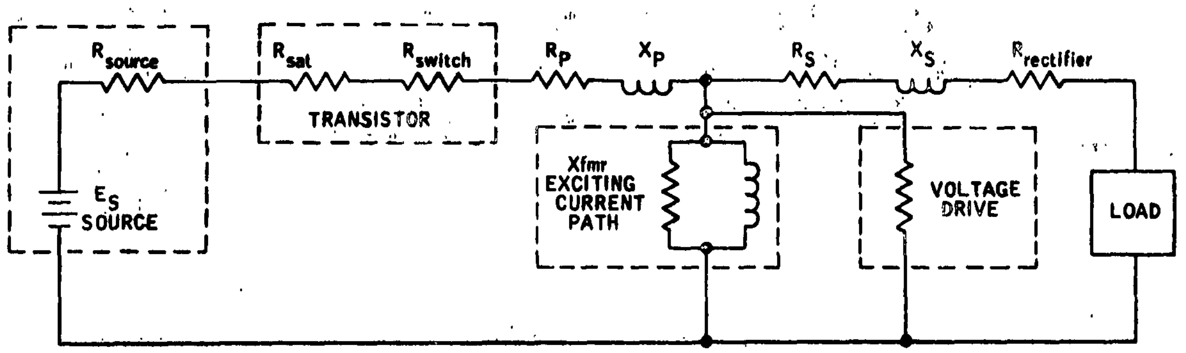


Figure 11 - EQUIVALENT CIRCUIT OF SELF-EXCITED CONVERTER HAVING VOLTAGE DRIVE

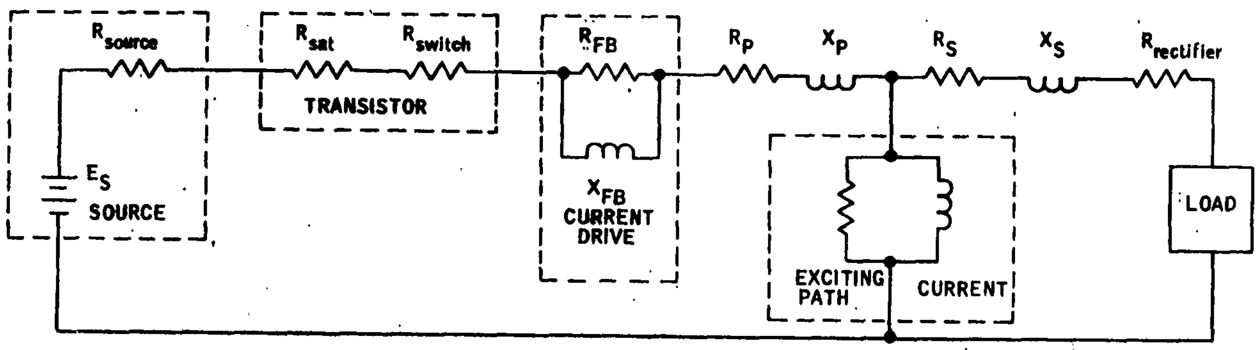
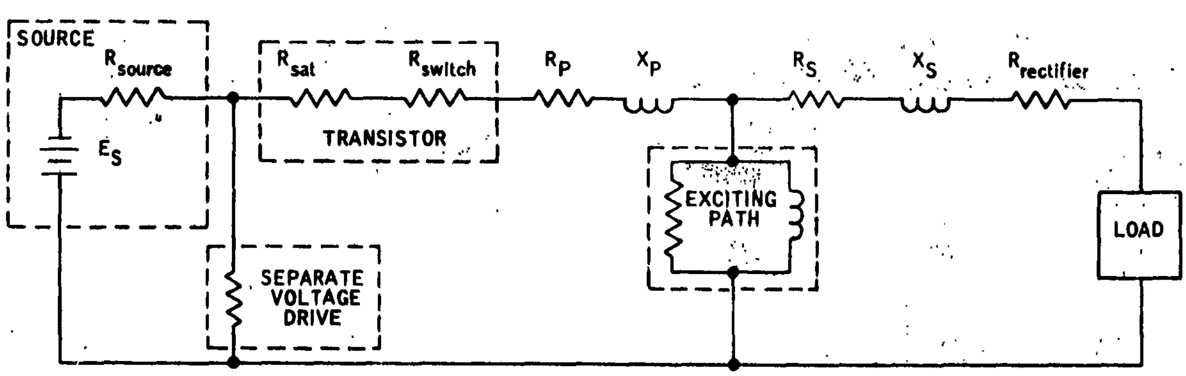


Figure 12 - EQUIVALENT CIRCUIT OF SELF-EXCITED CONVERTER HAVING CURRENT DRIVE

Figure 13 - EQUIVALENT CIRCUIT OF SEPARATELY EXCITED CONVERTER WITH SEPARATE VOLTAGE DRIVE



is a small voltage drop but is nevertheless an appreciable quantity, especially when the source voltage is low. The volt-amperes that remain after the current transformer drop is subtracted are consumed in the power transformer losses and are transformed to supply the load. Since transistor drive power required for both current and voltage feedback is the same for a given load, the circuit of Figure 12 should have basically the same efficiency as the circuit of Figure 11. The method of furnishing the drive is different but since the power requirements are the same and the losses are basically the same, the efficiency should be the same.

In actual operation, however, there will be some differences in the two circuits, especially if the load is varied over a wide range. The current feedback circuit will have drive power supplied proportionally to load, and hence at light loads, drive losses will be less. This circuit tends to provide a more optimum drive for all load conditions and operates near optimum efficiency at all loads. The voltage feedback converter will tend to operate at a more fixed drive and hence will have too much drive at light loads. This will result in a lower efficiency at light loads than can be obtained with a current feedback converter. Another factor affecting the voltage drive circuit is the high source impedance anticipated. At light loads, the effective source voltage will rise causing a proportionate increase in drive voltage. The drive power will rise at light loads causing additional losses which reduce light load efficiency. In addition, at extremely heavy loads the voltage drive may become insufficient. This analysis shows that the voltage feedback circuit must be designed for some optimum load, source voltage, and source impedance; whereas, the current feedback circuit can operate more efficiently over a considerable range of loads, source voltages, and source impedances.

Figure 13 shows the equivalent circuit of a separately excited converter having voltage drive. The transistor currents in this converter are slightly lower than in Figures 11 and 12 because the converter does not supply its own drive power. The slightly lower collector currents will tend to reduce transistor losses and increase the transistor gain slightly. This will enable the power amplifier section to operate at higher efficiency or have a greater power capacity. The separate exciter is a low power voltage feedback converter. The lower power requirements of the exciter will facilitate frequency control or synchronization at this point if desired. This converter will have to be designed for optimum performance at a fixed load since the voltage drive will not be optimum for all loads. The main advantage of the separate drive circuit is that the converter capacity can be slightly greater. A separately excited circuit may be easier to start. One disadvantage of separate excitation is the lack of inherent overload protection.

e. Other Circuit Arrangements - In order to minimize the series resistance reflected into the primary power circuit by the current transformer, an additional stage of amplification might be looked at. A circuit incorporating this feature is shown in Figure 14.

In this circuit, the current feedback supplies only the drive current for Q_3, Q_4 which are smaller transistors. The feedback transformer turns ratio is $(h_{FE1}) \dfrac{N_6}{N_5} (h_{FE_3})$. The reflected impedance is equal to

$$R_p' = R_{b_2} \left[\frac{1}{\left[\dfrac{(h_{FE_1}) N_6 (h_{FE_3})}{N_5} \right]^2} \right] \quad (1)$$

where:

h_{FE_1} = assumed gain of Q_1

h_{FE_3} = assumed gain of Q_3

N_5 = primary turns on T_3

N_6 = secondary turns on T_3

R_{b_2} = input impedance to Q_3

R_p' = reflected impedance in primary

It can be noted that the reflected impedance varies directly with R_{b_2} which is relatively high for a two ampere transistor and varies inversely with $(h_{FE_3})^2$. Some typical values would be R_{b_2} = 25 ohms

$$h_{FE_3} = 15$$

The incorporation of these values into equation (1) would give,

$$R_p' = \frac{25}{225}\left[\frac{1}{\left[\frac{h_{FE_1} \quad N_6}{N_5}\right]^2}\right] \quad \text{or}$$

$$R_p' = \frac{1}{9}\left[\frac{1}{\left[\frac{h_{FE_1} \quad N_6}{N_5}\right]^2}\right]$$

Thus it can be seen that some reduction in primary reflected impedance should be achieved by using this scheme. The actual drive power is obtained directly from the power source through the first stage amplifier. The output of this circuit would probably not produce drive proportional to load unless resistors were used in the emitters of Q_3, Q_4. This would tend to increase power dissipation and input impedance, defeating the purpose of reducing the reflected impedance in series with Q_1, Q_2. For this reason, the circuit does not appear to show any significant advantages. The increased complexity probably outweighs any advantages it may offer.

A modified Darlington type connection such as that shown in Figure 15 may also be considered. In this circuit the current feedback transformer is located in the primary and the current feedback is applied to the second transistor (Q_2, Q_4) in the Darlington pair. By using this arrangement, the feedback turns ratio can be increased considerably to reduce the reflected series impedance in the primary circuit. Assume the following conditions:

I_{c1} = 20 amps

I_{c2} = 2 amps

I_{b2} = .15 amp

E_1 = 3.0 volts

(R_b) Input impedance $= \frac{3.0}{.15} = 20$ ohms.

Feedback transformer turns ratio $= \frac{I_{b2}}{I_{c1}} = \frac{.15}{20} = .0075$

Thus 20 ohms is reflected into the primary by $(.0075)^2$.

The impedance reflected into the primary would then be:

$$R_b' = R_6 (.0075)^2 \tag{2}$$
$$= 20 \times .562 \times 10^{-4}$$
$$= 11.22 \times 10^{-4}$$
$$R_b' = .001122 \text{ ohms}$$

With I_{c1} = 20 amps. Voltage drop across current feedback primary = $I_{c1}R_b'$ or 20 amps x .001122 ohm = .02244 volt. This is somewhat lower than with current feedback to a single transistor, but is still an appreciable percentage of the saturation voltage drop. The percentage is $\frac{.0225}{.133}$ = 17 per cent.

The diodes CR1, CR2 provide back bias to turn Q_1 or Q_2 off rapidly. Resistors R_1, R_2 provide positive voltage feedback. This circuit is more complex than other drive circuits and any benefits obtained by this arrangement may not justify the increased complexity.

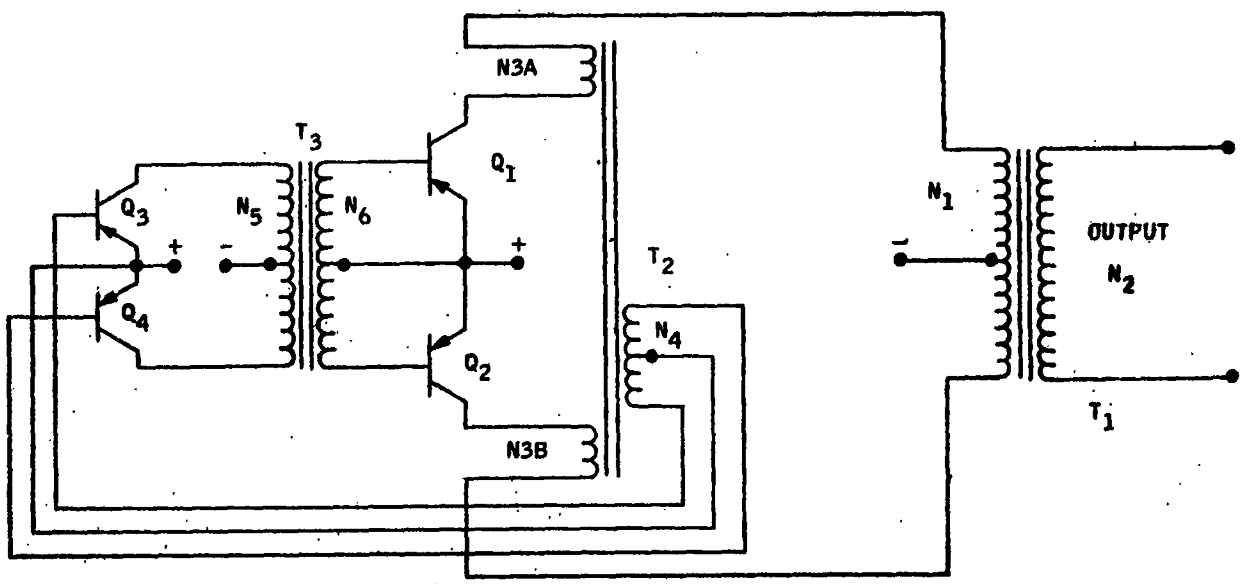


Figure 14 - CONVERTER WITH ADDITIONAL DRIVE AMPLIFICATION

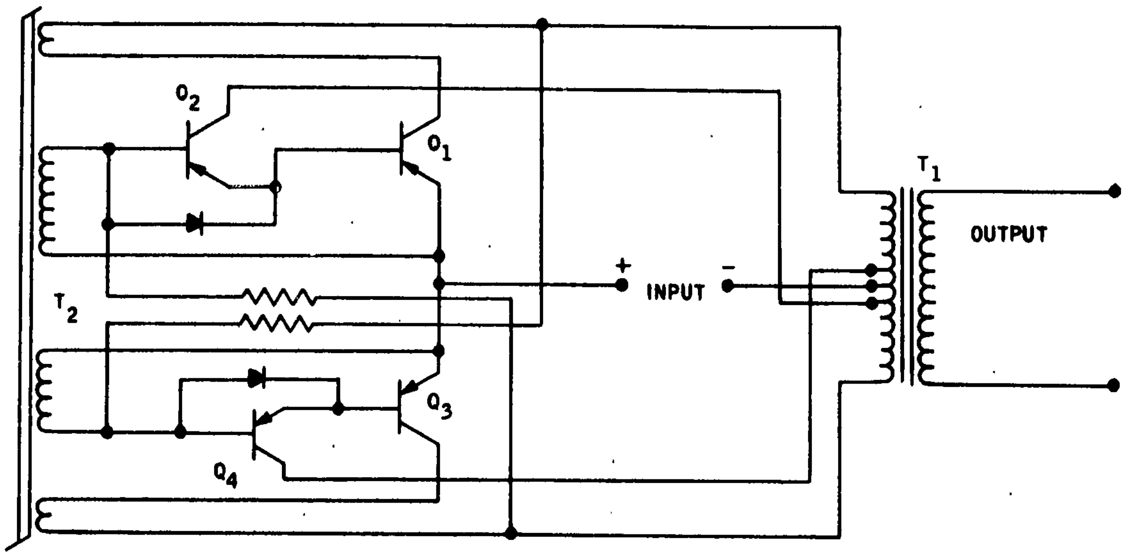


Figure 15 - CURRENT FEEDBACK WITH MODIFIED DARLINGTON CONNECTION

Some other circuits which may be considered to provide proportional current drive are shown in Figures 16 and 17. Figure 16 shows

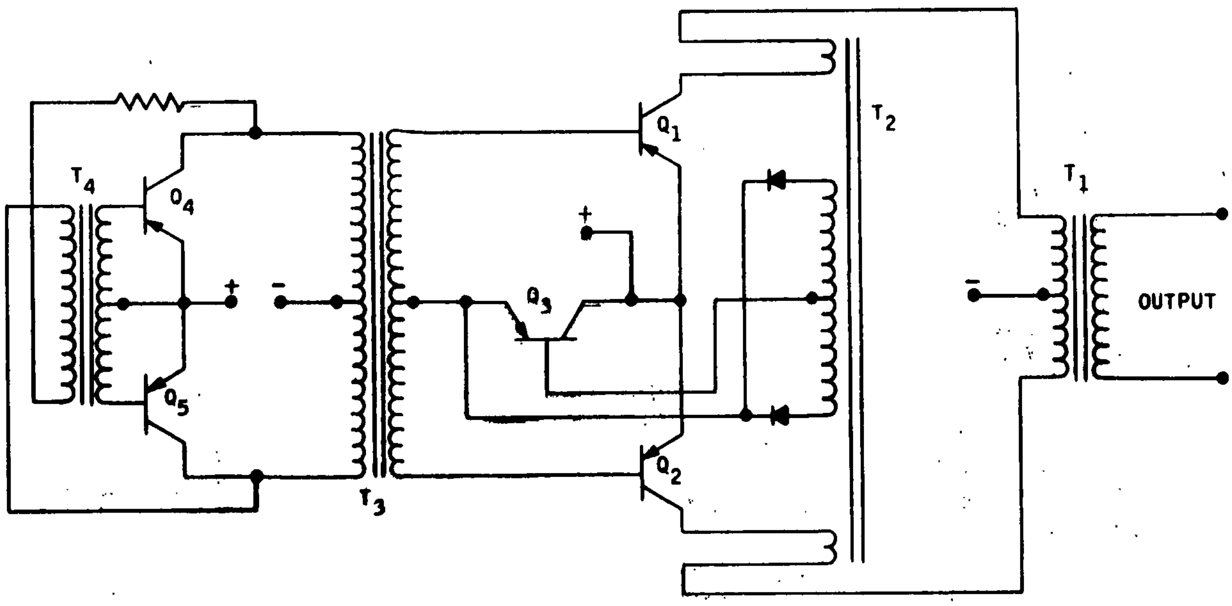


Figure 16 - CONVERTER WITH DRIVE CONTROL

a circuit in which the drive power is derived from a separate square wave inverter and regulated proportionally to load current by a series dropping regulator Q_3.

The high gain of Q_3 allows the control current to be low and hence a high turns ratio on T_2. Q_3 will have a relative high input impedance and hence the impedance reflected into the primary, although low, will still be significant. Also, the series dropping transistor Q_3 will contribute to power loss. This circuit contains additional control elements increasing the complexity. Because of these factors, this type of drive control circuit appears to have more disadvantages than advantages.

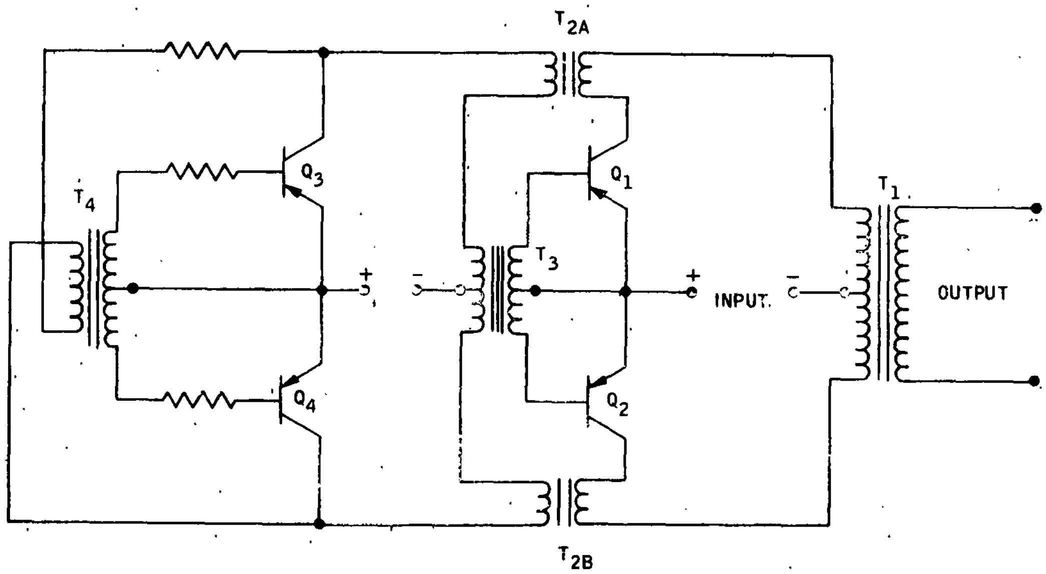


Figure 17 - SERIES SUMMED VOLTAGE AND CURRENT FEEDBACK

f. Combined Current and Voltage Drive - The circuit of Figure 17 combines voltage feedback from a separate exciter with current feedback from the power amplifier. In this circuit, the current feedback will add to the exciter supply voltage and provide additional drive power at heavy loads. Voltage and current drive power are summed in series with this arrangement. This circuit will provide a form of drive compensation for load variation but it will not be complete compensation.

Figure 18 shows a parallel connection of current feedback and voltage feedback.

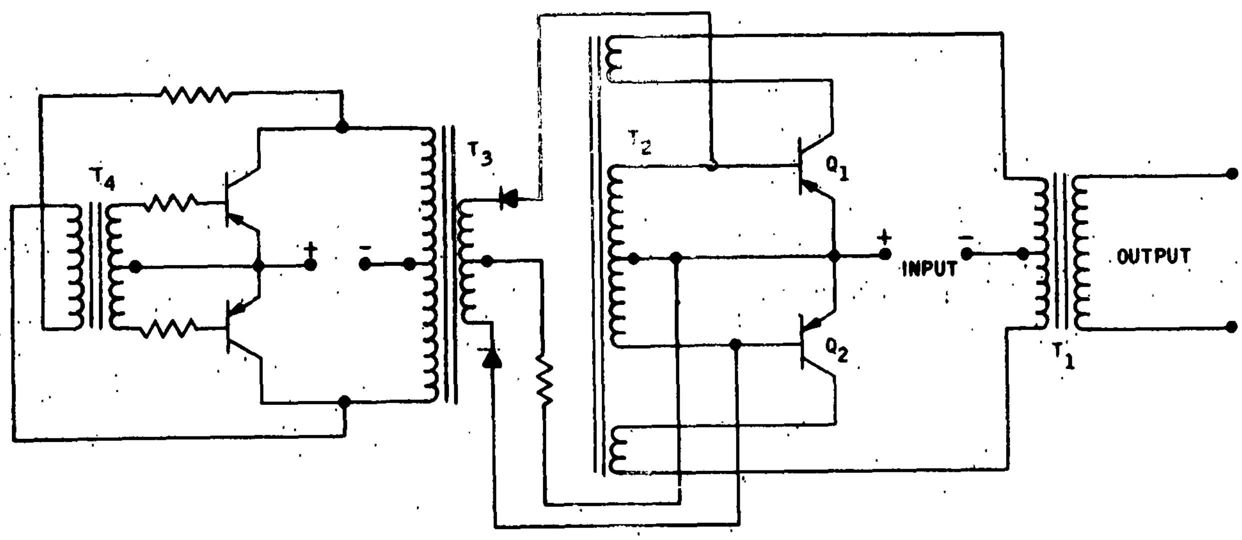


Figure 18 - PARALLEL SUMMED CURRENT AND VOLTAGE DRIVE

In this case the current feedback alone is sufficient to drive the transistors to the desired saturation level. With this degree of current feedback, adequate proportional drive will be furnished for all load and input voltage conditions. The function of the voltage feedback circuit is to synchronize the power amplifier to the required frequency and to supply over drive during the switching interval reducing switching time and switching losses. This type of circuitry has been used in higher voltage applications with wide source voltage and load variations and has been found to work quite well.* Because this circuit has desirable characteristics it will be considered for experimental verification.

3. Conclusions

The above investigation has shown the following:

1. Presently developed transistors do not have the required parameters for constructing a two transistor push-pull 50 watt converter which would operate at 75% efficiency from a 1.5 volt source.

2. Special transistors to meet the calculated parameter requirements might be developed if a lower resistivity and thinner base material is used and if advantage is taken of the low voltage requirements of most parameters.

3. The construction of a 50 watt converter operating from a 1.5 volt source with 75% efficiency may be presently feasible if the following is done:

   a) Use transistors from present production selected for low V_{CE} (Saturation) and high gain.

   b) Use four or more transistors in the converter power amplifier stage to reduce individual collector currents for operation at higher gains and lower saturation voltage drops.

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* Gemini-ACME Inverter DSG32A1 - Engineering Progress Letters 1 July - 1 August 1962; 1 August - 1 September 1962 Minneapolis-Honeywell Ordnance Division.

c) Select circuitry to provide optimum drive.

d) Select a feedback arrangement and adjust circuit parameters to provide rapid switching with minimum switching losses. Under (c) and (d) it is anticipated that a separate oscillator to supply voltage drive will be combined with current drive in the power amplifier to achieve an optimum drive. This device may consist of two converters operating out of phase so that one converter can supply all of the load while the other switches at light load to minimize transistor switching losses.

4. Twelve large power transistors would be necessary to construct a 150 watt converter operating from a 1.5 volt source at 75% efficiency.

5. The use of a combination of voltage and current drive is desirable to achieve maximum overall efficiency over a wide range of anticipated source voltage and load fluctuations.

6. The output transformer leakage parameters and the filter parameters must be optimized to provide the desired rapid switching.

7. The use of two converters with both primary circuits paralleled across the source and both secondary circuits paralleled across the load appears to provide the greatest reliability through redundancy. The use of two separate sources may also provide additional reliability.

8. Experimental verification of calculated results is desirable. This should be done by using a power amplifier with separate drive circuits. This method will provide the greatest flexibility and should provide information on the optimum switching speed and the drive required to obtain optimum results. A breadboard should be designed

so that current feedback and parallel power amplifier operation out of phase can be examined. Breadboard construction should be directed toward the recommendations of item 3 above.

9. The construction of transistor converters operating at 75% efficiency from voltage sources below 1.0 volt does not appear very favorable at this time. A considerable improvement in the transistor state of the art would be necessary to construct converters of appreciable power capacity at the required 75% efficiency.

Transistor Section prepared by:

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Project Engineer

C. TUNNEL DIODE APPROACH

Tunnel diodes have been considered as transducers for low input voltage converters because they have an extremely low forward saturation voltage drop. Typical commercially available germanium tunnel diodes have a forward drop (V_p) of 0.1 volt at 20 ampere peak currents*. Typical commercially available transistors carrying these current levels have a forward voltage drop of about 0.2 volt**. More optimum parameters might be obtained from selected units and selected operation points on both tunnel diodes and transistors. The tunnel diode saturation resistance is considerably less than on typical transistors. Thus the tunnel diode may be attractive because the saturation resistance loss is one of the major losses in low voltage transistor converters. Developmental tunnel diodes have higher current ratings*** (200 amps) than transistors (65 amps)**. These have a low 0.11 volt (V_p) at 200 amps which is lower than the 0.45 volt maximum V_{CE} (SAT) for 65 amp. transistors *** **.

1. Theoretical Efficiency of Tunnel Diode Converters

Our literature search shows that several papers have been written on tunnel diode converters and inverters and these show operating characteristics and theoretical equations for determining maximum operating efficiencies. The following papers have been examined:

1. Tunnel Diode D-C Power Converter

   Authors: H. F. Storm, D. P. Shattuck
   AIEE - Transactions - Communications and Electronics, July 1961

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* R. C. A, Tentative Data Sheet High Current Germanium Tunnel Diodes Developmental Type No. TD192.

** Honeywell Data Sheet MHT - 1803, 1903, 2003.

*** R. C. A. Tentative Data Sheet Extra High Current Tunnel Diodes: Type Nos. TD224, TD225, TD226.

2. Tunnel Diode Static Inverter

Author: J. M. Marzolf
Electrical Engineering, February 1962

3. Analysis of Tunnel-Diode Converter Performance

Author: D. J. Hanrahan
I. R. E. Transactions on Electron Devices, July 1962

Item No. 3 is of particular interest because equations are derived for the theoretical maximum efficiencies of the storm-shattuck d-c converter, a push-pull inverter, a push-pull converter, the Marzolf inverter, and the Marzolf converter. These equations are based on the assumption of zero losses in the transformers, rectifiers, capacitors, and wiring. One equation for the Marzolf circuit does consider magnetizing current but neglects other transformer losses. These equations establish the maximum efficiency limit on these tunnel diode converters. The equations show that tunnel diode oscillator efficiencies are limited by both current and voltage peak to valley ratios.

The typical characteristics of commercially available tunnel diodes were inserted into these equations to determine the upper efficiency limit that might be obtained with off-the-shelf tunnel diodes.

Reference 3 above shows that the basic efficiency equation for the storm-shattuck and push-pull converters is:

$$\eta_o = \frac{1}{\left(1 + \frac{2I_v}{\Delta I}\right)\left(1 + \frac{2V_p}{\Delta V}\right)} \qquad (3)$$

The efficiency of the Marzolf converter is:

$$\eta = \left[1 - \frac{I_m}{\Delta I} \right] \eta_o \tag{4}$$

where:

Referring to Figure 19,

η	=	overall theoretical efficiency
η_o	=	theoretical efficiency factor
I_v	=	valley current
I_p	=	peak current
V_v	=	valley voltage
V_p	=	voltage drop at peak current
ΔI	=	$I_p - I_v$
ΔV	=	$V_v - V_p$
I_m	=	magnetizing current

The following are typical values for a commercially available (*) 200 ampere tunnel diode:

Peak current to valley current ratio, $(I_p/I_v) = 8:1$

I_p = 200 amps.

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* R. C. A. Tentative Data Sheet Extra High Current Germanium Tunnel Diodes Developmental Type No. TD226

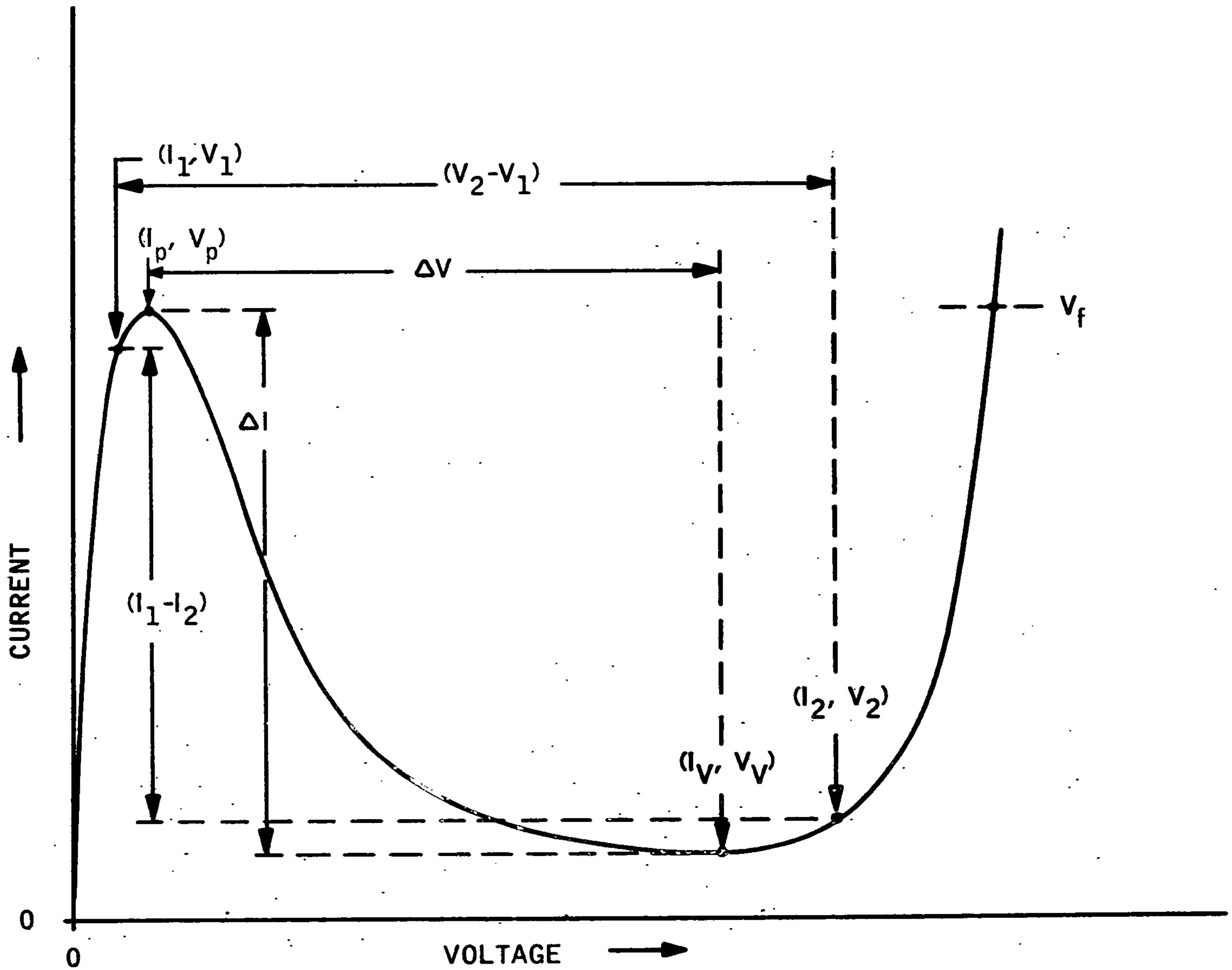

Figure 19 - TUNNEL DIODE CHARACTERISTICS AND NOMENCLATURE

Voltage at peak current, V_p = .11 volt maximum

Valley voltage, V_v = .370 volt maximum

Forward blocking voltage at I_p, Vf = .480 volt maximum

These values give:

$$I_v = \frac{200}{8} = 25 \text{ amps}$$

$$\Delta I = I_p - I_v = (200-25) = 175 \text{ amps}$$

$$\Delta V = (V_v - V_p) = (.37 - .11) = .26 \text{ volt}$$

From this data η_o can be determined:

$$\eta_o = \frac{1}{\left(1 + \frac{2I_v}{\Delta I}\right)\left(1 + \frac{2V_p}{\Delta V}\right)} \tag{3}$$

$$= \frac{1}{\left(1 + \frac{2(25)}{175}\right)\left(1 + \frac{2(.11)}{.26}\right)} = \frac{1}{(1.286)(1.846)} = \frac{1}{2.374}$$

$$\eta_o = 42.2\ \%$$

From these results it can be seen that the maximum efficiency obtainable with a typical commercially available device is about 42% (assuming no other losses) and this is much less than the 75% efficiency desired. Much improvement is necessary to achieve the desired results. If characteristics were taken from specially selected tunnel diodes rather than typical state of the art devices the theoretical efficiency may be somewhat higher but would probably not exceed 60%. Thus improvement is necessary to achieve the desired efficiency.

Because of the low limit for theoretical maximum efficiency with a typical state of the art device, it is desired to determine what the characteristics of a tunnel diode should be to construct a device having an efficiency of 75%. From the above calculations it can be noted that η_o could be broken into two factors: $\frac{1}{(1 + \frac{2I_v}{\Delta I})}$ and $\frac{1}{(1 + \frac{2V_p}{\Delta V})}$. The typical values gave $(\frac{1}{1.286})$ or .78 for the former and $(\frac{1}{1.846})$ or .542 for the latter. It is apparent that the greatest improvement could be made by reducing the ratio of $\frac{2V_p}{\Delta V}$.

We shall assume that a Marzolf converter is used and the theoretical efficiency neglecting transformer, wiring, diode, and filter losses is:

$$\eta = \left[1 - \frac{I_m}{\Delta V}\right]\left[\frac{1}{1 + \frac{2\,I_v}{\Delta I}}\right]\left[\frac{1}{1 + \frac{2V_p}{\Delta V}}\right]$$

The factor $\left[1 - \frac{I_m}{\Delta I}\right]$ considers the effects of transformer magnetizing current. This includes the core loss but does not include primary and secondary copper loss. If an assumed value of transformer efficiency is substituted for this quantity a more accurate approximation may be obtained. In addition to transformer losses there will also be switching losses and rectifier and filter losses. Thus to consider these losses, η_o should be multiplied by the efficiencies of these other factors. The use of practical operating points should also be considered.

The above equations are valid for any operating point since they do not consider losses incurred during switching but merely give the theoretical efficiency for operation at the quiescent "on" and "off" operating points. Operation at the threshold peak and valley points may be somewhat difficult since the frequency would vary considerably and there may also be a tendency to break into a much higher frequency oscillation.* For these reasons, calculations on desired tunnel diode characteristics use optimistic practical operating points and efficiency estimates for the transformer, rectifiers, and switching losses.

2. Calculated Tunnel Diode Requirements

Calculations on tunnel diode characteristics necessary to achieve 75% and 65% overall efficiency are shown in Appendix B. The calculated parameter requirements are shown in Tables IV, V, VI, VII, and VIII in Appendix B. Curves have been plotted in Figures 20, 21, and 22 showing the calculated tunnel diode characteristics. The curve of Figure 20 shows the desired characteristics for a tunnel diode necessary to build a 50 watt converter operating from a 0.25 volt source at 75% efficiency. This curve has a peak current to valley current ratio of 35.1.

The current operating point ratio is 28.7. It has a valley to peak voltage ratio of 10 and a voltage operating point ratio of 15.77. Note that the operating point ratios have been shifted from peak values to more optimum values which reduce the current ratio and increase the voltage ratio. The low 0.25 volt input necessitates a low 0.042 volt peak voltage. Examination of published specification sheets indicates that peak voltages, V_p, normally range from 0.07 to 0.13 volt.

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* Temperature and input voltage variations may cause parameter variations which will move the operating points from the threshold region.

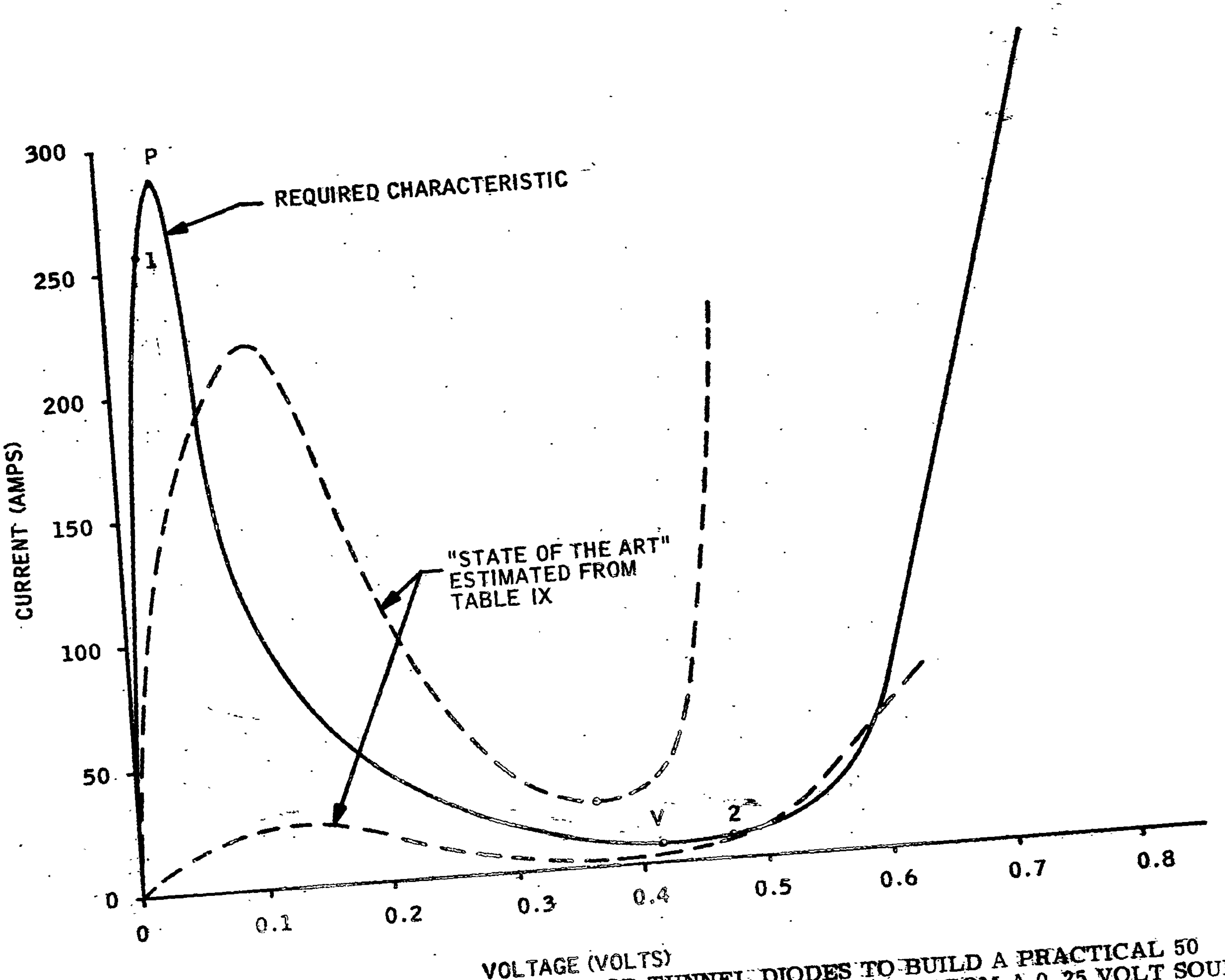

Figure 20 - CALCULATED REQUIREMENTS FOR TUNNEL DIODES TO BUILD A PRACTICAL 50 WATT dc-dc CONVERTER OPERATING AT 75% EFFICIENCY FROM A 0. 25 VOLT SOURCE

With respect to the peak voltage, V_p, energy band considerations indicate that V_p should be equal to the Fermi level degeneracy, assuming a symmetrical junction. It might seem that this degeneracy and V_p could be made arbitrarily small, but this conflicts with the need for high tunnel current density. Basically, all the voltage characteristics of the tunnel diode are functions of the material and its impurity concentrations and therefore should be independent of the current magnitudes and physical size.

It appears unreasonable to expect a V_p value as low as 0.042 volt in a high current device using germanium, silicon, or gallium arsenide materials. It is reasonable to expect about 80 millivolts. Since the ratio of $\frac{V_p}{(V_v - V_p)}$ is one of the important factors which determines efficiency, consideration has been given to the use of tunnel diodes made from materials having wider band gaps and higher valley voltages such as silicon and gallium arsenide. Calculations have therefore been directed toward this and are located in Appendix B.

The calculated tunnel diode requirements necessary to construct 50 watt converters operating at 75% efficiency from 0.515 volt and 0.614 volt sources are shown in Figure 21. These characteristics are based upon a peak voltage of 0.08 volt which has been assumed reasonable. As in the characteristics described above, the operating point has been chosen at 90% of the peak current and at 110% of the valley current. The use of a higher peak voltage, V_p, has resulted in higher valley voltage and higher source voltage requirements. These two curves have valley voltages at 0.88 and 1.08 volts. Preliminary investigation has not revealed any high current tunnel diodes with valley voltages this high. The use of the higher source voltage does reduce the peak current requirement I_p to about 138 and 115 amperes for the 0.515 and 0.614 volt inputs respectively. It is assumed that the operating points chosen are close to optimum. One of the factors which should be considered in a more accurate analysis is the determination of optimum operating points on a given tunnel diode characteristic curve. To do this, the equation of the curve is desired. Calculations have been made to determine tunnel diode

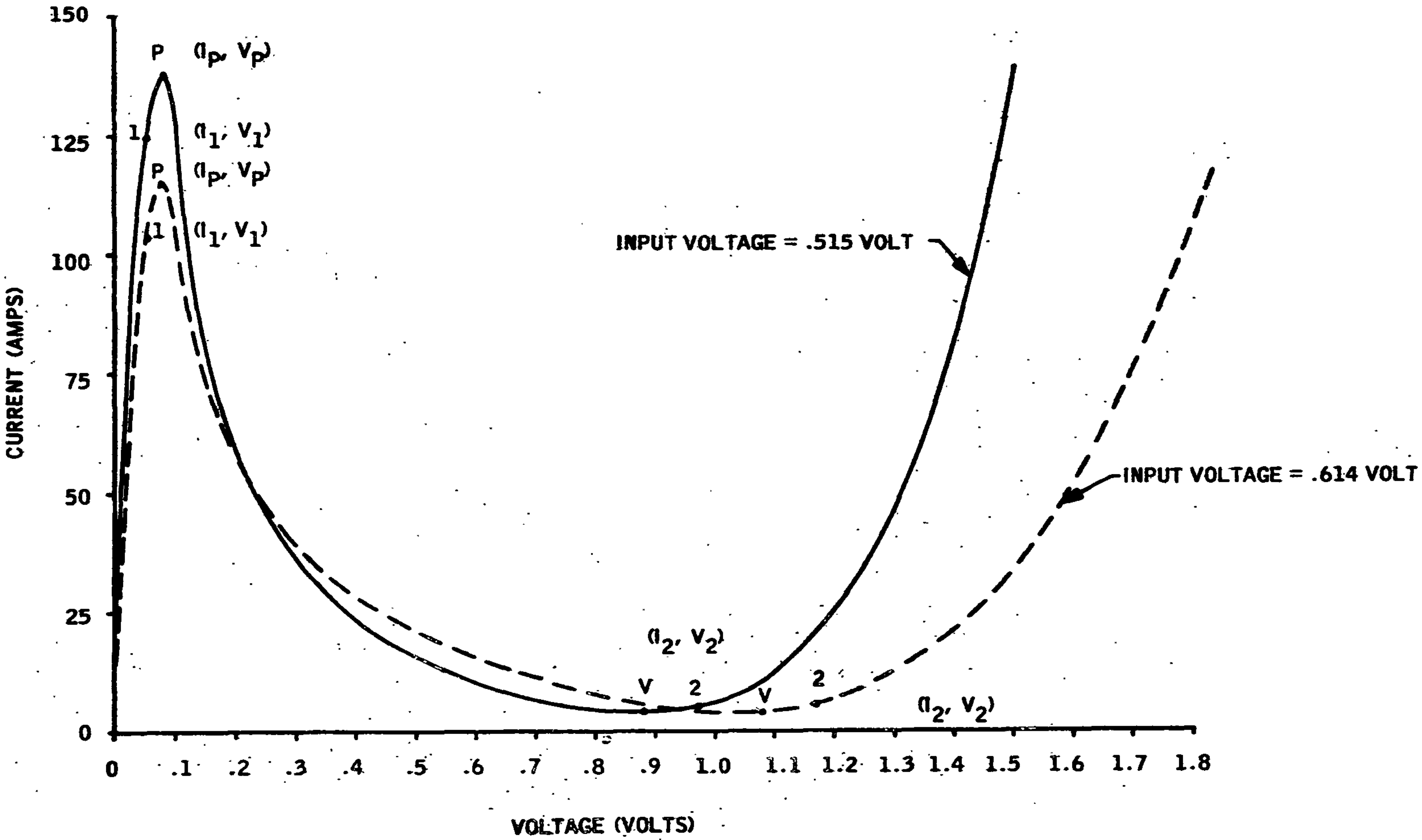

Figure 21 - CALCULATED REQUIREMENTS FOR TUNNEL DIODE CHARACTERISTICS NECESSARY TO BUILD A PRACTICAL 50 WATT CONVERTER HAVING 75% EFFICIENCY

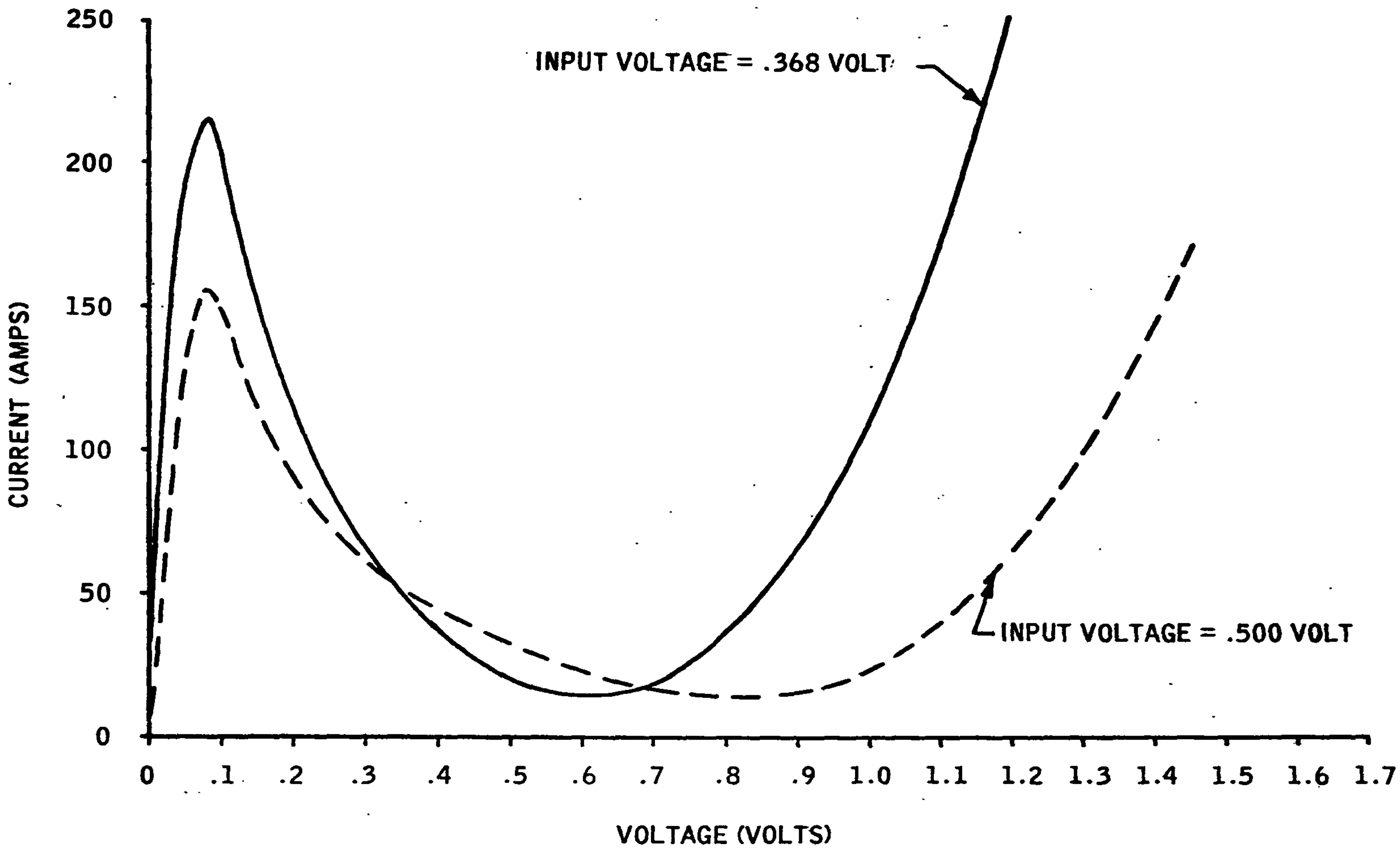

Figure 22 - CALCULATED REQUIREMENTS FOR TUNNEL DIODE CHARACTERISTICS NECESSARY TO BUILD A PRACTICAL 50 WATT CONVERTER HAVING 65% EFFICIENCY

characteristics necessary to build 50 watt converters operating at 65% efficiency from 0.368 and 0.500 volt sources. The calculations are in Tables VII and VIII, Appendix B, and the curves are shown in Figure 22. The curve for the .368 volt input has a peak voltage of 0.08 volt and a valley voltage of 0.587 volt. These voltage points begin to approach the realm of possibility. The peak current is 214 amperes and the valley current is 14.7 amperes. The current and voltage ratios are 14.5 and 7.34 respectively. These ratios are still high but closer to commercially available characteristics than the 75% requirements. To date, devices which meet all of these requirements have not been found. The curve for the 0.50 volt input has a peak current of 154.1 amps and a valley current of 13.6 amps. The valley voltage, however, is 0.85 volt which may be difficult to obtain.

3. Tunnel Diode State of the Art

Table IX in Appendix B shows typical parameters for presently available germanium, silicon, and gallium arsenide tunnel diodes. By comparing the data for available commercial units in Table IX with the requirements of Tables IV, V, VI, VII and VIII, and curves of Figures 20, 21, and 22, it can be seen that considerable improvement is necessary. It is anticipated that improvements will be made in these ratios, and selected units may have better characteristics. Although much higher ratios have been reported it is not known if higher current ratios and higher voltage ratios have both been obtained simultaneously in the same research device. An exceptionally high current ratio may have been obtained at the expense of other parameters such as the voltage ratio. It is known that several manufacturers are working on higher current devices which may satisfy our peak current requirements. High current tunnel diodes appear entirely feasible. The problem lies in improving the peak to valley ratios.

Personnel familiar with the tunnel diode state of the art were contacted and the following opinions on future development possibilities were obtained:

1. Silicon tunnel diode peak current to valley current ratios are about five. This is caused by high valley currents. Because the current ratio is low with only moderate voltage ratios, silicon devices do not appear very favorable for high efficiency applications.

2. The state of the art of germanium devices is the most advanced. Since much work has been done on germanium there are no significant material problems. It may be possible to push the germanium device peak voltage down to 50 millivolts. However, if this is done the valley voltage tends to decline also. For germanium material, voltage ratios up to six and current ratios up to twelve might be obtained. With these limitations, the theoretical maximum efficiency for germanium tunnel diode converters would be about 60%.

3. Gallium arsenide is the more unknown semiconductor material. The peak current to valley current ratio for this material may be as high as 20. The peak voltage is about 125 millivolts. The voltage ratio of present gallium arsenide devices is about 4.5. Gallium arsenide may be the best material but considerable work must be done to obtain higher voltage ratios. Much more effort would be required to develop high current gallium arsenide devices than germanium because effort must be expended on material improvement as well as device development. It is not known how rewarding effort exerted on developing gallium arsenide devices would be.

4. Valley voltages for any of these materials will probably be limited to less than .6 volt.

Our preliminary survey shows that high current germanium tunnel diodes will probably remain the best choice for the tunnel diode approach in the near future. This means that the tunnel diode approach will probably be limited to an overall efficiency of 50% to 60% in the immediate future. Other losses may reduce the efficiencies of practical converters to slightly less than 50%.

4. Conclusions

Calculations have specified the tunnel diode parameters necessary to achieve 75% and 65% efficiencies for 50 watt converters. The required tunnel diode parameters are beyond the present state of the art. High current tunnel diodes are available and can be developed but the required peak current to valley current and valley voltage to peak voltage ratios cannot be obtained at the present time. The primary limitation on overall efficiency is caused by the low valley voltage to peak voltage ratio. For this approach, high current germanium tunnel diodes appear to be best for the immediate future. Gallium arsenide may eventually provide superior tunnel diodes but the development of suitable high current devices is not foreseen in the immediate future.

Fifty watt germanium tunnel diode converters will have theoretical upper efficiency limits between 50% and 60%. Practical converters will be limited to slightly less than 50% due to other circuit losses.

The source voltage must be held within narrow limits for maximum tunnel diode converter efficiency. Thus the input voltage and power source types are limited by the tunnel diode characteristics. To operate from a higher source voltage it may be possible to use two tunnel diodes in series. This of course would not affect the theoretical efficiency which is a function of parameter ratios. Some low voltage sources such as thermoelectric generators have high internal impedance which will cause high output voltage fluctuations with load variations. The output voltage of such a power source will vary considerably due to temperature differential changes at the hot and cold junctions during warm-up, operation, and subjection to various environments. Operation of a tunnel diode converter from a power source with this type of wide voltage fluctuations would be extremely difficult. These wide voltage fluctuations might cause high dissipation and might cause operation in the positive resistance region preventing oscillation. Tunnel diode converter operation from a source with large voltage variations does not appear promising.

It can be concluded that this approach does not appear favorable for 75% efficiency requirements. If the efficiency requirements can be lowered to 50% or if new developments occur, then this approach should be reevaluated.

Tunnel diode section prepared by:

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* The author wishes to thank Mr. J. T. Maupin for assistance and consultation in preparing the tunnel diode section of this report.

D. ELECTROMECHANICAL APPROACH

During this quarter, feasibility of a particular electromechanical approach configuration was investigated. This approach, shown in Figure 23, consisted of an oscillating "U" tube liquid metal chopper driven by a Faraday type pump. The device operates as follows:

An alternating magnetic field (B) and a high current d-c field (I) are impressed across the liquid metal duct perpendicular to each other producing a force on the fluid according to Fleming's left hand rule. The alternating flux field causes the force field to vibrate producing vertical oscillations in the fluid, opening and closing contact with electrodes A and B. The amount of excursion, x, must be sufficient to provide the following:

1. Close the contact with sufficient over-travel to provide low contact resistance and prevent contact bounce due to ripples in the liquid surface.

2. Open the contact sufficiently so that surface tension effects will be overcome and the electrode and fluid will part.

3. Maintain sufficient open gap to prevent arcing and to prevent contact bounce due to surface ripples.

These requirements definitely place limitations on the minimum excursion that can be used.

In order to operate satisfactorily, the liquid metal conductor and pressurized gas must be maintained in separation. If the two were to mix, the conductive fluid column would rise in the tube and both contacts would be closed at the same time producing a dead short through A, B, and C. Separation of the two media is dependent primarily upon the earth's gravitational field. Surface tension also has some effect, but this is relatively small compared to the gravitational field.

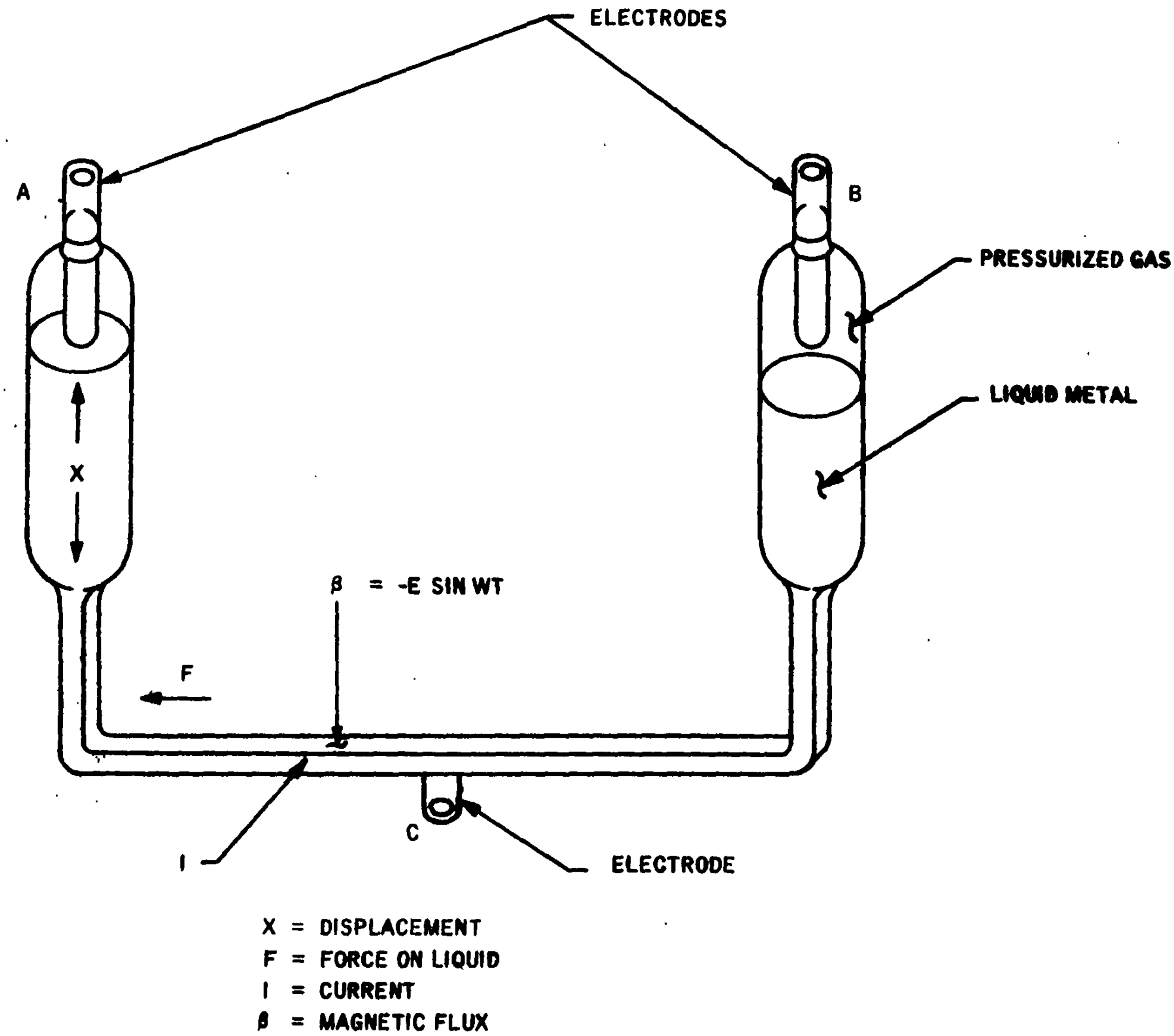

FIGURE 23 - OSCILLATING "U" TUBE CHOPPER SKETCH

An analogy of the liquid in the tube is made in Figure 24.

In this analogy, the fluid is assumed to be in a beaker fastened to a piston in a cylinder wall. The simple harmonic motion is applied to the fluid via the crank, connecting rod, piston, and beaker. Examination of this figure shows that equations can be written for the displacement, velocity, and acceleration of the fluid.

The displacement x can be found from:

$$x = R \sin w t \tag{5}$$

where:

x = displacement of fluid

R = length of crank

w = angular velocity of crank in radians

t = time

The velocity of the fluid can be found by differentiating equation (5):

$$v = \frac{d\,x}{d\,t} = R\,w \cos w\,t \quad \text{or}$$

$$v = R\,w \cos w\,t. \tag{6}$$

The acceleration of the fluid can be found by differentiating equation (6):

$$a = \frac{d\,v}{d\,t} = R\,w^2\,(-\sin w\,t) \quad \text{or}$$

$$a = -R\,w^2 \sin w\,t. \tag{7}$$

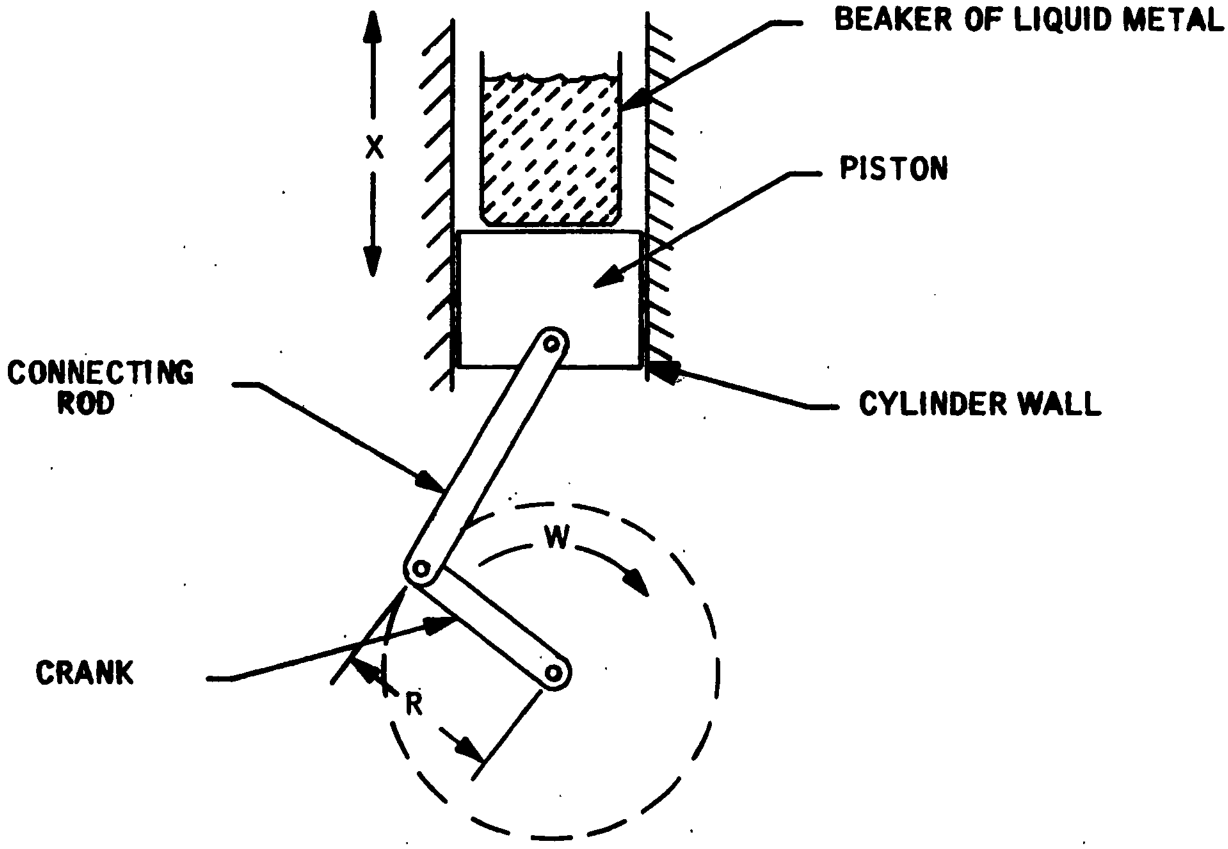

Figure 24 - OSCILLATING FLUID COLUMN ANALOGY

The force on the fluid will be determined by the sum of the acceleration fields upon it. Thus:

$$F = m\,(a + g) \text{ or} \tag{8}$$

$$F = m\,(g - Rw^2 \sin wt) \tag{9}$$

where:

F	=	force on fluid
m	=	mass of fluid
a	=	acceleration due to vibratory excursion
g	=	acceleration due to gravity = 980 cm/sec.^{2}

It can be noted that ($-Rw^2 \sin wt$) will subtract from the gravitational field whenever ($\sin wt$) is positive. It is therefore obvious that if $|RW^2|$ equals the earth's gravitational field, the net force on the fluid will be zero and a state of weightlessness will be obtained. If $|RW^2| > g$ then the fluid will fly out of the beaker. In the actual device of Figure 23, it would fly into the gas section and close the contacts causing a short.

This analysis shows that a critical point exists which will cause failure when:

$$Rw^2 \sin wt \geq g \tag{10}$$

Actually there will be secondary surface ripples in this device which will also add to this condition. These will be generated when the fluid contacts and separates from the electrode. It is unlikely that the secondary oscillations would be quickly damped out. The secondary surface oscillations would also have a simple harmonic motion acceleration.

$$a_s = R_s\, W_s^{\,2} \sin W_s\, t_s \tag{11}$$

where: the subscript s denotes the secondary effect. A more accurate formula for critical operation would be when:

$$R w^2 \sin wt + Rs\, W_s^{\,2} \sin (W_s + \phi)\, t \geq g \tag{12}$$

where: ϕ varies with surface position.

It would be reasonable to assume that some point on the fluid surface will have a maximum secondary acceleration when the main column is at the maximum acceleration point. This analysis shows that equation (10) is optimistic in that the secondary oscillations are not considered.

1. Sample Calculation

By neglecting secondary surface ripples and surface tension effects, it can be seen that the most critical time in equation (10) will occur when sin wt = 1. Under this condition, equation (10) reduces to:

$$R w^2 \geq g \text{ or} \tag{13}$$

$$w \geq \sqrt{\frac{g}{R}} \tag{14}$$

The value of g is a constant 980 cm/sec.^{2} The value of R can be assumed with practical limits of contact resistance, contact bounce, and contact clearance in mind. An excursion of 0.2 cm appears sufficiently optimistic. This would result in an R of 0.1 cm.

Using the above values, the critical angular velocity w can be determined.

$$w_c \geq \sqrt{\frac{g}{R}} \quad \text{where } w_c = \text{critical value.} \tag{14}$$

$$w_c \geq \sqrt{\frac{980 \text{ cm/sec.}^2}{0.1 \text{ cm.}}} \geq \sqrt{\frac{9800}{\text{sec}^2}}$$

$$w_c \geq 99 \text{ radians/sec.}$$

One cycle equals 2π radians. The maximum critical frequency f_c can be found from:

$$f_c = \frac{W_c}{2\pi} = \frac{99 \text{ radians/sec.}}{2\pi \text{ radians/cycle}} \tag{15}$$

f_c = 15.75 cycles per second. A curve of critical frequency is plotted vs. R and is shown on Figure 25. Note that critical frequencies above 30 cycles per second require extremely small excursions which would not be practical when contact clearance requirements are considered.

2. Conclusions

Preliminary investigation of this particular configuration shows that a theoretical critical maximum operating frequency exists. The sample calculation shows a maximum critical operating frequency of 15.75 cycles/sec. for a device with an optimistic total excursion of only 2 millimeters. In Figure 25, which shows critical frequency vs. total excursion, it can be seen that maximum operating frequencies are very low when practical distances are considered for the open contacts. These results indicate that the maximum operating frequency could not be higher than 30 cycles per second. In actual practice, the frequency would probably have to be lower than this because the secondary ripples and surface tension effects would tend to provide additional problems.

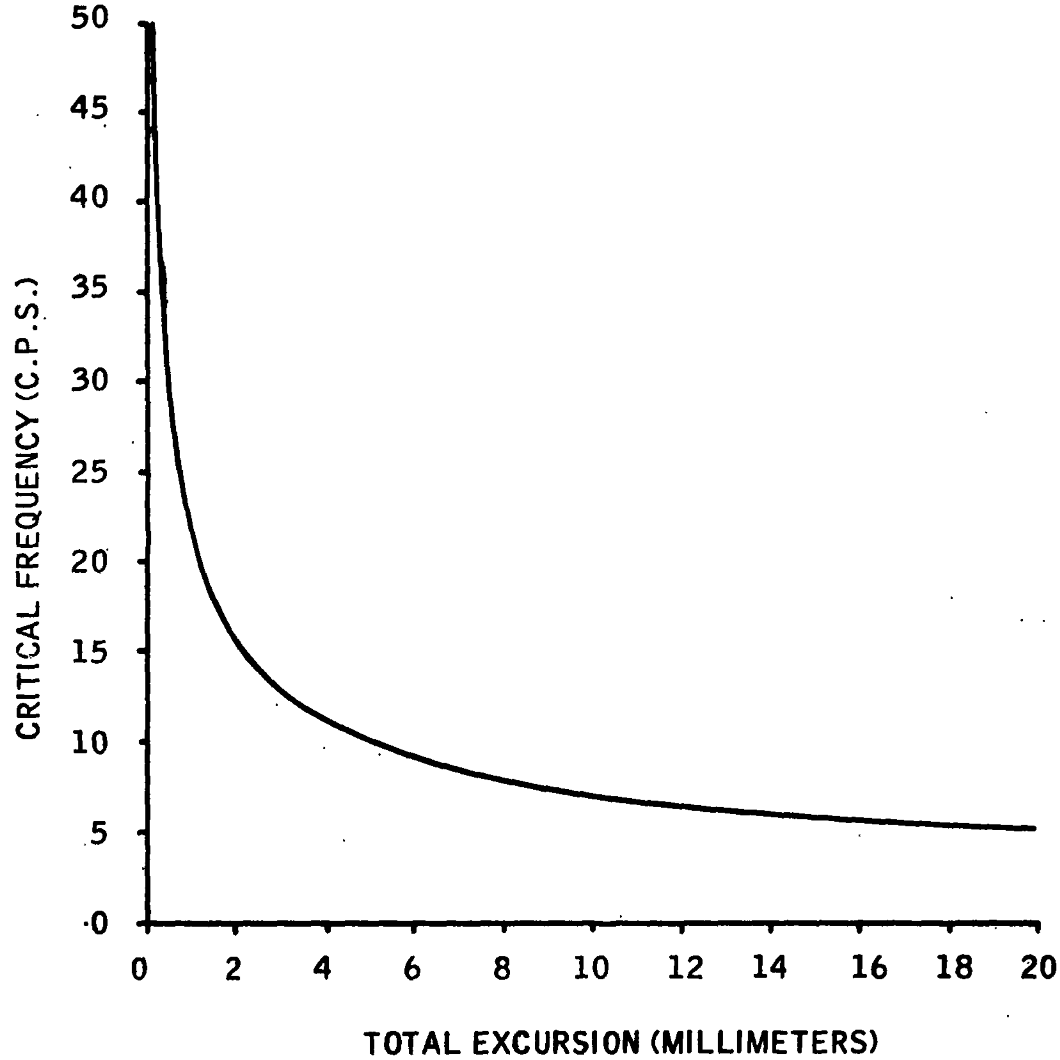

Figure 25 - CRITICAL FREQUENCY VS. TOTAL EXCURSION

The present form of this approach depends upon the force of gravity to maintain separation between the liquid metal conductor and the pressurized gas insulator. Analysis indicates that when the acceleration of the fluid equals the acceleration force due to gravity, the conductive fluid will be in an uncontrolled weightless state tending to prevent controlled contact opening. This would result in the closing of both contacts and hence a dead short at frequencies higher than the critical frequency. The critical frequency varies inversely with the square root of the amplitude of vibration.

This low maximum operating frequency would result in a heavy and impractical device for field use. The dependence upon the gravitational field for satisfactory operation is the cause of the frequency limitation. As long as the gravitational field is a prime factor in this device this approach will not appear very favorable.

There may be other configurations or modifications which are not as severely limited by gravitational fields. Further study and investigation should be made on the electromechanical approach to discover more optimum configurations to examine. Some approaches which might be considered are the wetted contact relay type, (relay contacts wetted by liquid metal held in position by capillary action), and solid contact type.

Electromechanical section prepared by:

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E. LIQUID METAL MAGNETOHYDRODYNAMIC CONVERTERS

One possible method of converting extremely low input voltages to higher voltages consists basically of a liquid Metal Magnetohydrodynamic Pump which circulates liquid metal through a Magnetohydrodynamic Generator. A simple illustration of such a device is shown on Figure 26. As shown in the figure, a magnetic field and a current field are applied to the liquid metal channel all perpendicular to each other forming a liquid metal pump. By Fleming's left hand motor rule the liquid metal is a current-carrying conductor in a magnetic field and a force is applied perpendicular to the two fields causing the fluid to flow.

The fluid is pumped to a separate generator section having its own magnetic field, contact surfaces, and different channel dimensions. The conductive fluid flowing through the generator channel induces an output voltage by Fleming's right hand generator rule. The generator voltage can be stepped up by increasing fluid velocity, magnetic strength, or length of the current path through the channel.

The magnetohydrodynamic relationships involved in this device are quite complex. Honeywell has been working in this field for several years and a Design Manual has been written for designing devices of this type.*

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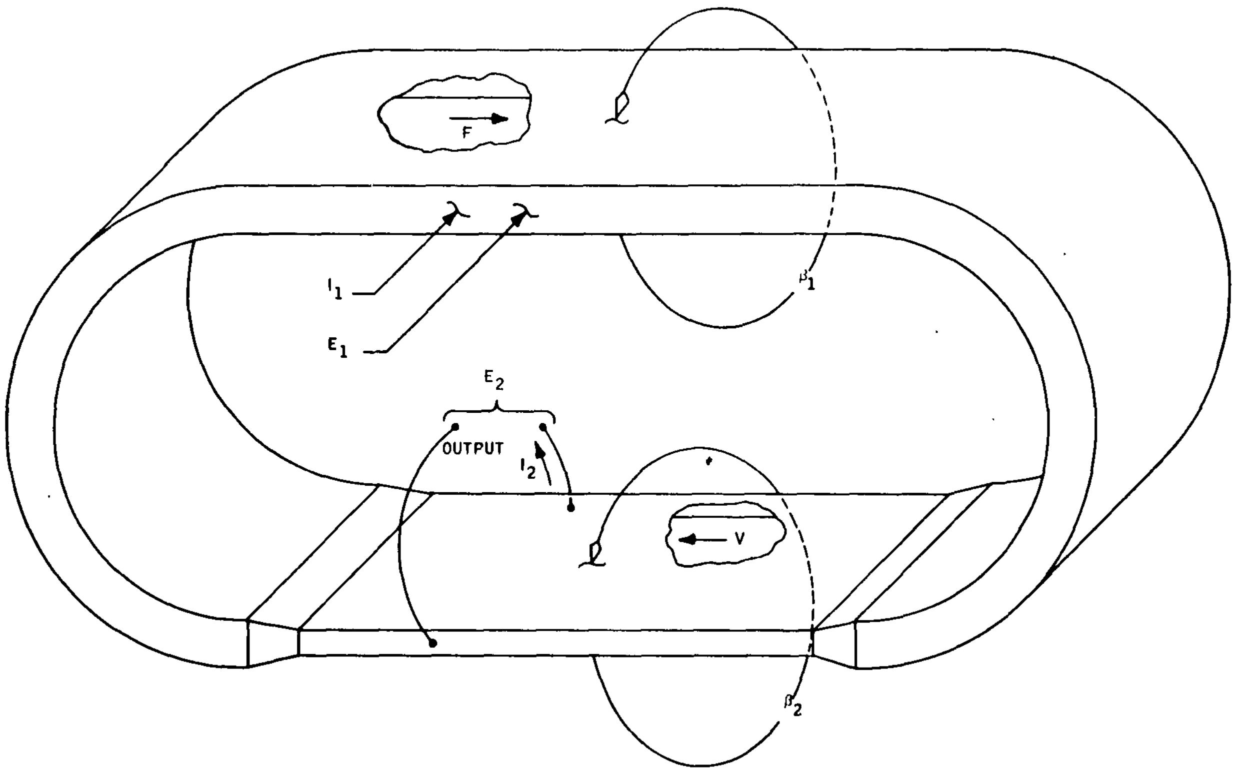
* Fluid Armature Electric Motors and Generators -- Design Principles HR-60-476 by Frank M. Exner, Minneapolis-Honeywell Research Center.



E_1 = PUMP INPUT VOLTAGE
I_1 = PUMP INPUT CURRENT
β_1 = PUMP MAGNETIC FIELD
F = FORCE CAUSING FLUID FLOW
V = FLUID VELOCITY
β_2 = GENERATOR MAGNETIC FIELD
E_2 = GENERATED VOLTAGE
I_2 = GENERATED CURRENT

Figure 26 - SIMPLE LIQUID METAL FARADAY MAGNETOHYDRODYNAMIC DEVICE

During this quarter Dr. J. E. Anderson reviewed the magnetohydrodynamic pump literature and checked the method of analysis of fluid flow in the magnetohydrodynamic channel. Dr J. E. Anderson concluded that the method of analysis used in the Honeywell Design manual was correct.

As part of Honeywell's Research program in fuel cell controls, a Magnetohydrodynamic Modulator preliminary model has been designed and fabricated. This device was constructed to verify the equations and design principles in the above design manual *. The above work was not part of this contract. Because of the Honeywell Research Center's current effort in this area, no extensive effort has been made on this approach under this contract during the first quarter. Our effort on this approach to date has been directed at obtaining information on their progress, reviewing the literature, and organizing our proposed study of this approach. More effort was not expended under this contract because it appeared desirable to wait for the results from the above program before investigating this approach. Waiting for these results has two advantages:

1. The information and experience gained from this model will be available at no cost to this program.

2. Present contract funds will be conserved since some of the work and verification desired will have been accomplished.

3. The experience gained from effort on the first model should result in an investigation of a more optimum configuration for this program.

Tests on the above model have been completed but evaluation and analysis of the results is still in progress. Some difficulty was encountered with the model and the reasons are known. These can be considered and avoided in future studies.

The most significant results of the above program have been the verification of some design principle assumptions and formulas in our Design Manual. (*)

It is anticipated that investigation of the liquid metal magnetohydrodynamic approach will commence in the second quarter. Both direct and modulating configurations will be considered. The initial effort may be directed toward direct d-c to d-c conversion because it is simpler and rectifier losses are eliminated.

Liquid Metal Magnetohydrodynamic Section prepared by:

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(*) Fluid Armature Electric Motors and Generators--Design Principles HR-60-476 by Frank M. Exner, Minneapolis-Honeywell Research Center.

F. TRANSDUCER BASIC EFFICIENCY CALCULATION METHODS

Converters using magnetoresistive, superconductive, photoresistive, and other approaches may be reduced to a push-pull equivalent circuit utilizing a DPDT switch. This equivalent circuit, Figure 27, switches "high" and "low" resistors in series with each half of the push-pull output transformer primary in order to chop the dc to ac for transformation. The formulas for efficiency and reflected load impedance have been derived. The formula for optimum load to achieve maximum efficiency has been derived. The formula derivations are shown in Appendix C. Calculations have been made for transducer impedance ratios from 1:1 to 2,000:1 at the optimum load condition. Curves have been plotted on Figure 28 showing Basic Efficiency, Optimum Load Resistance Referenced to Primary, and a defined Optimum Synthetic Resistor Rf all vs. Transducer Impedance Ratio over the above ratio range. The above formulas and Basic Circuit are shown in Appendix C and on Figure 28.

These calculations consider only the quiescent operating condition when one transducer is conducting at its lowest resistance and the other is at the high resistance state. Thus switching losses have not been calculated. These may differ with the various approaches. It is desired to consider the transformer, rectifier and filter, and switching efficiencies separately as lumped parameters. Using this method, the required basic efficiency figure must be raised above 80% in order to achieve a device having a calculated efficiency of 75%.

1. Conclusions

Preliminary examination of these calculated results indicates that the "off" transducer to "on" transducer impedance ratio should exceed 400 if construction of converters with 75% efficiency is to be within the realm of possibility. These

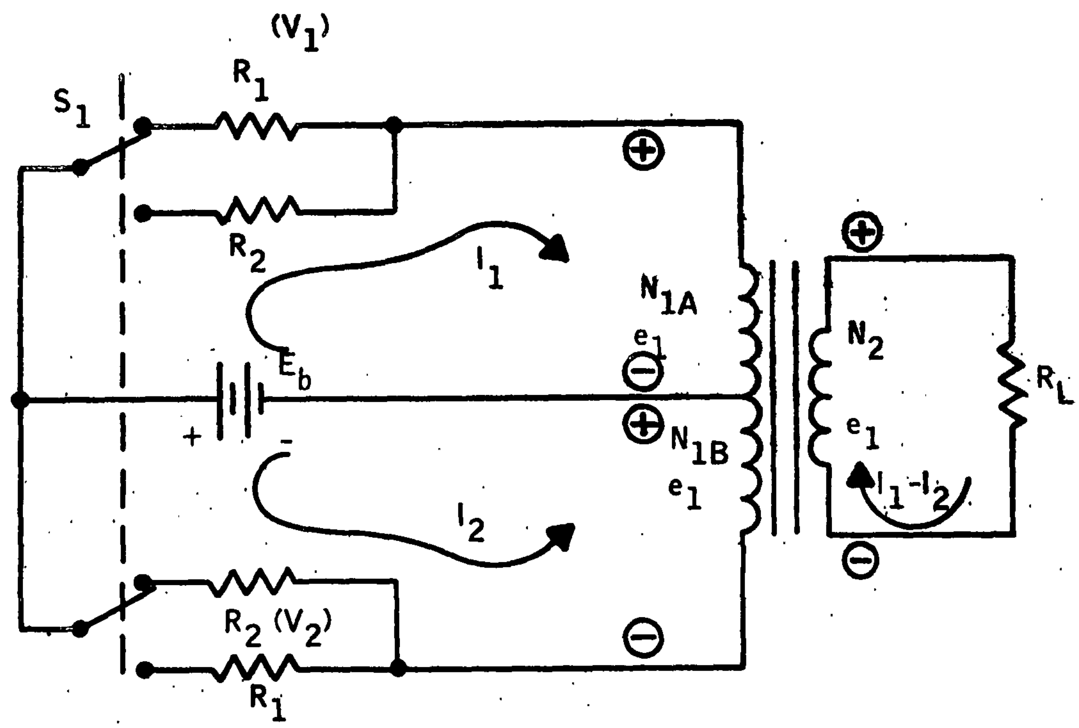

Figure 27 - PUSH-PULL EQUIVALENT TRANSDUCER CIRCUIT

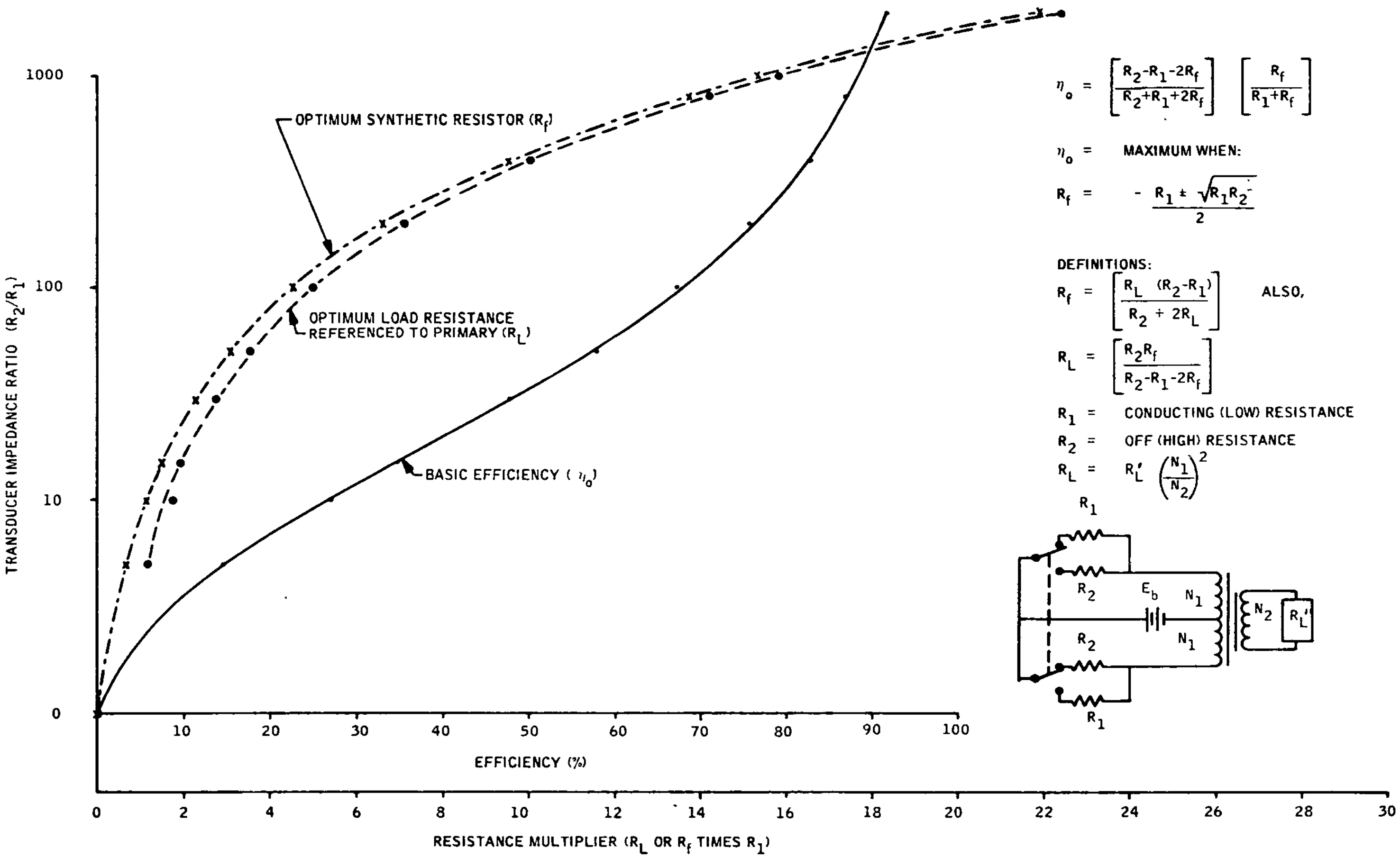

Figure 28 - CALCULATED PERFORMANCE OF PUSH-PULL CIRCUITS INCORPORATING SWITCHING RESISTIVE TRANSDUCERS

results have been useful in determining the feasibility of constructing the required transducers and converters using magnetoresistive, superconductive, and photoconductive effects.

Basic Efficiency Calculations prepared by:

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Project Engineer

G. HALL EFFECT AND MAGNETORESISTANCE CONTROLLED CONVERTERS

The Hall and Magnetoresistance effects in semiconductors provide possibilities for the low voltage conversion objectives of this contract. The purpose of this section is to evaluate these possibilities in terms of the required voltage conversion specifications and how they might be met with feasible Hall or magnetoresistance devices.

Our goal is a device which will convert 0.1 to 1.5 volts dc with an output power of 5 to 150 watts to a higher d-c voltage which is compatible with the voltage requirements of military electronic equipment. A Hall or magnetoresistance device could be used as a switch in a chopper amplifier to achieve the desired specifications.

1. Hall Effect

The Hall effect consists of a transverse voltage V_H developed in the z-direction of a rectangularly shaped sample of a conductor or semiconductor when a magnetic field H is applied in the y-direction perpendicular to a current I flowing in the x-direction through the sample. The situation is illustrated in Figure 29. We will consider the potentialities and restrictions of the Hall effect as applies here.

A Hall effect device is really a voltage converter, in the sense that the magnetic field converts the applied voltage V_A (which provides the current flowing through the sample in the x-direction) into an output Hall voltage V_H. However, this direct sort of conversion is of no value to us, because the ratio $V_H:V_A$ can never be even as large as unity, let alone greater. This is because the Hall voltage arises from the rotation of the lines of equipotential in the sample in a magnetic field. Values of $V_H:V_A$ only as high as about 0.7 have been observed. Thus, there is a fundamental limitation in the Hall effect which requires that $V_H < V_A$ always.

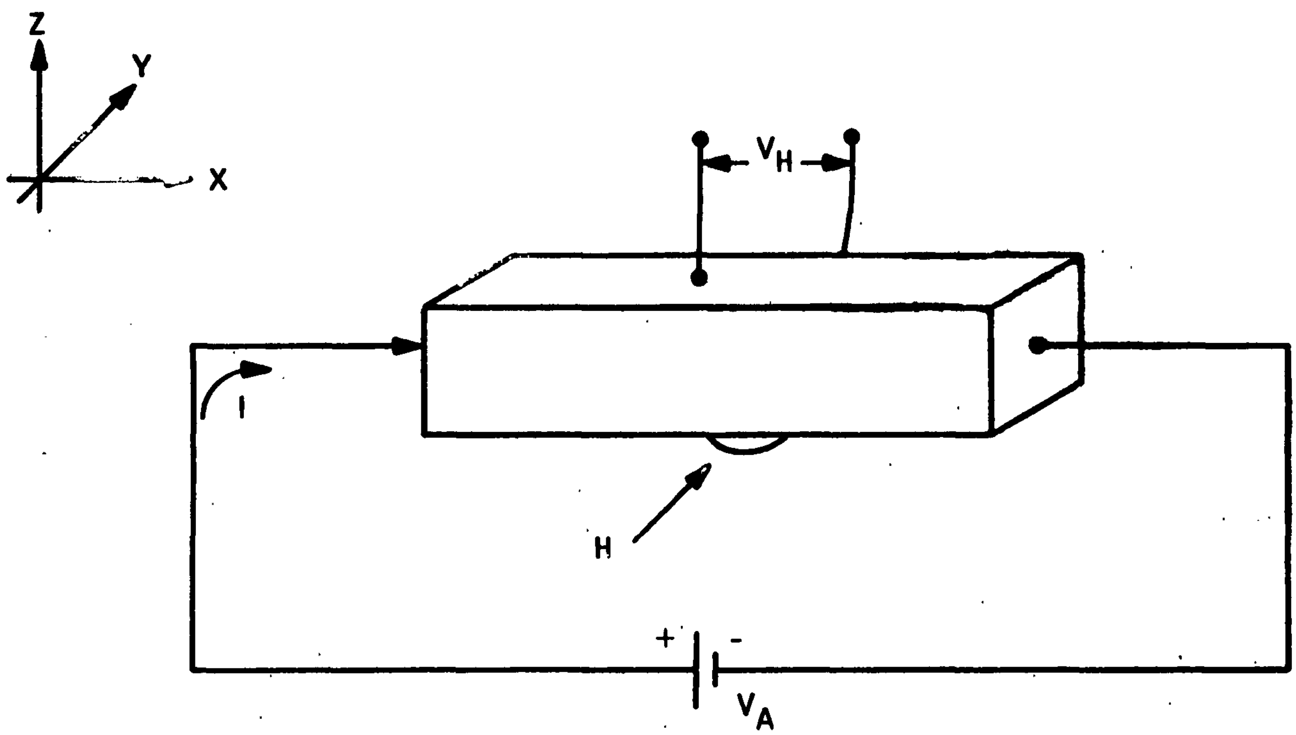

Figure 29 - HALL EFFECT

Another important point is that the power conversion efficiency of the Hall device will be too low. In fact, it is possible to argue that the maximum attainable efficiency is only about 17%, even under the most ideal conditions. Several papers on this subject contain detailed and rigorous mathematical calculations of maximum attainable efficiency. * Thus, if the Hall effect were used with an a-c magnetic field in order to convert d-c to a-c for subsequent voltage amplification, 83% or more of the power from the source would be lost in the Hall device, and this is a fundamental limitation. This limitation cannot be overcome, even with the use of better materials or any other improvement. Therefore, the Hall effect appears to be unpromising for our purposes and should not be considered further.

2. Magnetoresistance

The magnetoresistance effect is simpler to describe than the Hall effect, but its origin is more difficult to understand. The effect consists of a change in the electrical resistance of a sample of conductor or semiconductor when it is placed in a magnetic field.

This resistance change is always an increase, with a few unimportant exceptions. The detailed dependence of the resistance change on the magnetic field strength H varies in different materials and even in the same material at different temperatures. A common form of the dependence would be

$$R_H = R_o \ (1 + AH^n) \ , \tag{16}$$

where R_H is the field-on resistance, R_o is the zero-field resistance, A is a constant, and n is in the vicinity of 1 to 2. Note that we always have $R_H > R_o$.

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* See "Indium Antimonide as a Fluxmeter Material", E. W. Saker, F. A. Cunnell, and J. T. Edmond, British Journal of Applied Physics, Vol. 6, p. 217, June 1955.

The magnetoresistance effect occurs essentially because the magnetic field deviates the current carriers from their normal straight paths of drift through a sample, thereby making it "harder" for them to carry a given amount of current and thus increasing the sample resistance. In order for a material to exhibit a large magnetoresistance effect, the carrier mobility in the material must be high. The carrier mobility is defined as the drift velocity per unit applied electric field and expresses the ease with which a material carries an electric current. For a given applied voltage, given sample dimensions, and a given number of current carriers (electrons or holes), a sample will carry more current at higher mobility.

The material exhibiting the highest carrier mobility and therefore the largest known magnetoresistance effect is indium antimonide (InSb). The remainder of this section will discuss the potentialities and limitations of InSb "magneto-resistors" for the voltage converter application. In this discussion, we shall examine the situation in which the magnetic field is directed perpendicular to the sample current, which gives the largest effect.

An ordinary filament or rectangularly-shaped sample of InSb will have its current flow along the long dimension. In such a sample, there will also be a Hall effect, even if there are no Hall probes to measure it, since the magnetic field orientation which produces the magnetoresistance inevitably gives a Hall effect. This Hall effect and its electric field are undesirable for our purposes, because the Hall field tends to keep current carriers drifting along straight paths, instead of being deviated by the magnetic field as much as possible in order to get a large magnetoresistance effect. The Hall field can be minimized, and the current deviations promoted, by making the length-to-width ratio of the sample small. In other words, it is desirable to use a short "stubby" sample of the type shown in Figure 30. The current contacts in the

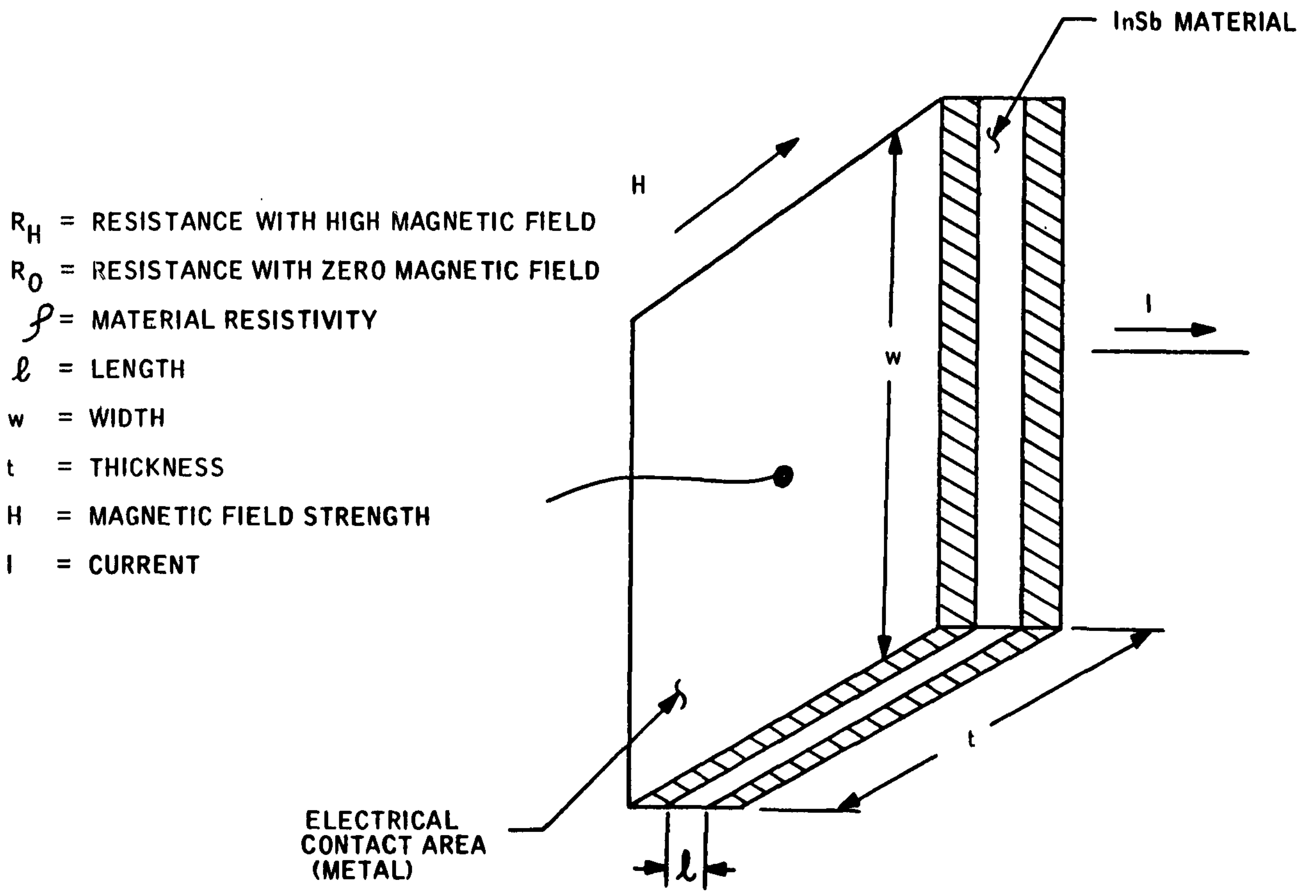

Figure 30 - STUBBY MAGNETORESISTOR

sample in Figure 30 serve to short out the Hall voltage to a great extent. Studies of this sample-shaping approach have been made by several people* in the past, and it was found that a considerable enhancement of the magnetoresistance effect was achieved. An even greater enhancement can be achieved by using a Corbino disk shaped InSb sample, as illustrated in Figure 31.

Let us consider a rectangular short stubby sample shape like that of Figure 30, and return later to the Corbino disk. The resistivity ρ of a pure sample of InSb at around room temperature ($\sim 70°$) is approximately 6×10^{-3} ohm-cm. For our purposes, the magnetoresistor should have as low an R_o resistance as possible, since it is intended for use as a switch in converting voltages by a chopper amplifier type of technique. The particular R_o will be determined by several considerations, including the magnet design and magnetoresistor fabrication problems, so we will defer a more quantitative discussion until later. For the sample shape of Figure 30, we will have

$$R_o = \rho \, \ell / wt \, , \tag{17}$$

where the symbols are defined in Figure 30. This type of sample has given magnetoresistance ratios as high as

$$\frac{R_H}{R_o} \cong 20 \tag{18}$$

at temperatures around 70° F in a magnetic field of 10,000 gauss.

For a Corbino disk-shaped sample, the zero-field resistance is given by

$$R_o = \frac{\rho}{2\pi t} \ln \frac{r_2}{r_1} \, , \tag{19}$$

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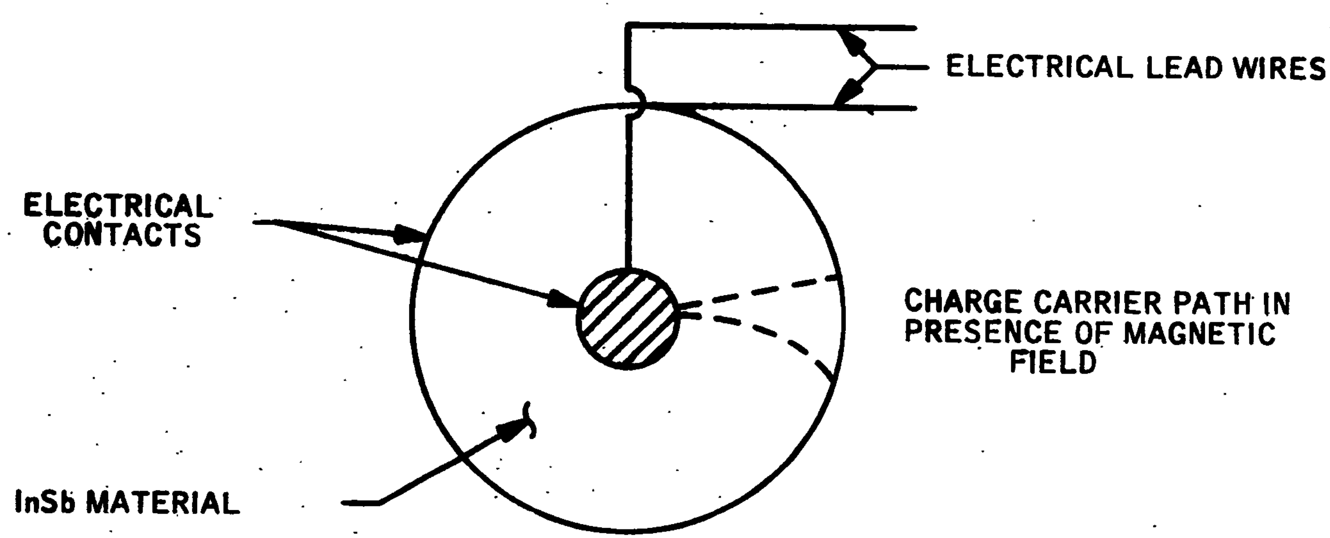
*H. Weiss, Journal of Applied Physics, Supplement to Vol. 32, p. 2064, 1961. A. C. Beer, ibid, p. 2107, 1961.



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Outer radius is – r_2
Inner contact radius is – r_1
Thickness is – t

Figure 31 - CORBINO DISK

where the symbols are defined in Figure 31. Values of the magnetoresistance ratio of as high as

$$\frac{R_H}{R_o} \cong 30 \tag{20}$$

have been observed at 10,000 gauss around room temperature in Corbino disks of InSb.

The magnetoresistance ratios can be made greater by using the InSb magnetoresistor at lower temperatures, but the resistivity will be somewhat larger than the room temperature value quoted above. For example, a very pure (n-type) InSb sample of the type which would give the largest possible magnetoresistance would have a resistivity at liquid nitrogen temperature of about 4×10^{-2} ohm-cm, which is almost 10 times the room temperature resistivity. Thus, if the limitation on the feasibility of an InSb magnetoresistor is in the attainable low R_o value, it would not help to decrease the temperature. On the other hand, if the main limitation is the attainable R_H:R_o ratio and a rather high R_o can be tolerated, one might achieve ratios of over 100 for R_H:R_o at 77°K in a 10,000 gauss field. In fact, values of nearly 300 for R_H:R_o have recently been reported for very pure samples of InSb. These data were taken on Corbino disk samples. Let us discuss the Corbino disk potentialities in more specific terms.

Since the requirements for the InSb sample to be used as the magnetoresistive element are:

a) Low resistance
b) Large change in resistance in a magnetic field
c) a geometry suitable for use in a small magnet gap,

a Corbino disk type sample, shown in Figure 31, is proposed. Since the resistance of a Corbino disk depends on the ln r_2/r_1 (Equation (19), where r_2 and r_1 are defined in Figure 31, the sample resistance can be made very small by using a ring shaped sample in which r_1 is only slightly less than r_2. The magnetoresistance effect will be the largest with a Corbino disk geometry, as the Hall voltage is not present. Since the current flow is radial, as shown in Figure 31, the current carriers, when deflected by the magnetic field, are not deflected to a sample surface where they accumulate to form a Hall voltage as in a conventional sample but are constantly deflected by the magnetic field so that they spiral from the inner electrode to the outer electrode. This mechanism results in the very large magnetoresistance effect in materials such as InSb having a high carrier mobility. Since the magnetic field is perpendicular to the plane of the disk, a small magnet gap may be used by making the Corbino disk very thin (small t).

A d-c to a-c converter utilizing the magnetoresistance effect would consist of the basic circuit shown in Figure 27. This is directly analogous to a circuit with transistors or mechanical switches. The magnetoresistance elements would be switched from their low resistance to high resistance states by the application of a magnetic field. This changing magnetic field could be supplied by an electromagnet, a rotating permanent magnet, or by moving the sample in one of these magnets. The magnet current would be provided by the "stepped-up" d-c of the converter; thus, this is a loss to be deducted from the over-all efficiency.

The efficiency of the circuit shown in Figure 27 is determined primarily by the ratio of the "on" to the "off" resistance of the semiconductor magneto-resistance element. A plot of efficiency as a function of the ratio of the two resistances has been shown in Figure 28. It has been shown previously that the Corbino disk structure produces the greatest magnetoresistance change. The resistance of a Corbino disk structure has been shown previously to be:

$$R = \frac{\rho}{2\pi t} \ln \frac{r_2}{r_1}$$

Since the converter is to operate from low voltage high current sources, the zero field resistance must be very low. This would imply directly that the thickness t must be large; however, that would make the magnetic field structure correspondingly larger. Therefore, the ratio of outside to inside radius $r_2:r_1$ must be small. As a result of the above considerations, it can be shown that a thin ring is the best shape for low resistance and high magnetoresistance.

Initial calculations indicate that an element with an outside diameter of one centimeter, inside diameter of 0.8 cm, and a thickness of 0.2 cm yields a resistance of 10^{-3} ohms and an "on" to "off" resistance ratio of about 20 at room temperature. This gives a maximum conversion efficiency of 40%. An electromagnet to produce the required 10,000 gauss field would be the size of a cube three inches on each side and would dissipate four watts of power. In a converter system operating with 150 watts input, 45 watts would be dissipated in each magnetoresistance element. This makes it obvious that careful consideration must be given to cooling the elements. As pointed out, the magnetoresistance decreases considerably with increasing temperature. One possible magnet and element configuration is shown in Figure 32. To provide good electrical and thermal contact with the element, the InSb might be cast directly in the heat shield as a polycrystalline material. This would greatly ease the problem of handling the brittle InSb without substantially reducing the magnetoresistance, provided that the casting is not badly polycrystalline.

Much could be gained by operating the entire magnet and magnetoresistance element in liquid nitrogen. First, the magnetoresistance would increase from 20 to about 200, thus raising the basic efficiency from 40% to 75% Second, it would reduce the loss in the magnet due to the electrical resistance of the copper. This would permit a much smaller magnet. Third, the problem of cooling would be considerably reduced because of the good thermal transfer to the liquid nitrogen. For a system operating with an input of 150 watts at 75% efficiency, 0.89 quart of liquid nitrogen would be vaporized per hour.

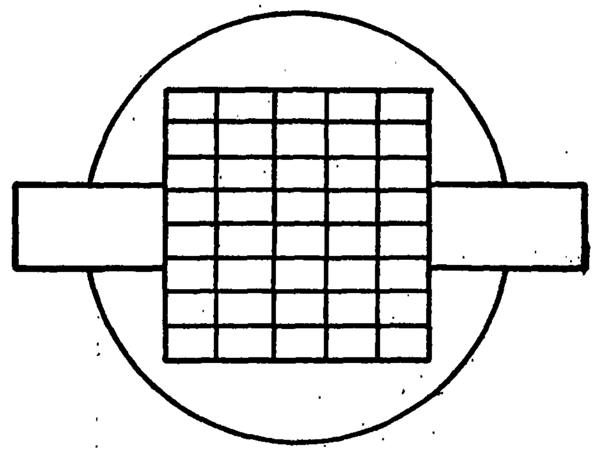
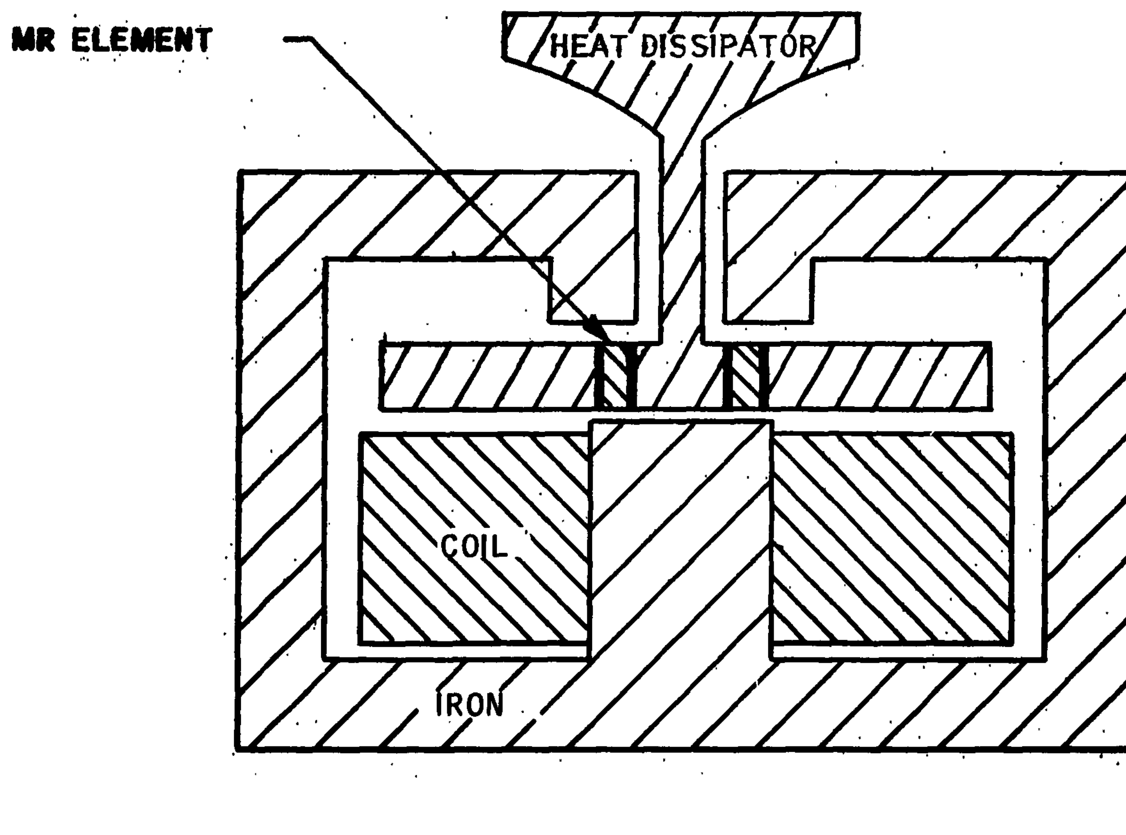


Figure 32 - A POSSIBLE MAGNET AND MAGNETORESISTOR CONFIGURATION



3. Conclusions

Further consideration should be given now to the engineering problems involved in the design of a practical magnetoresistive converter. It is not yet clear that this type of converter is feasible, mainly because it is not certain that the heat dissipation problem can be solved.

It has been concluded that the Hall effect approach is not feasible for this application. Further work must be done on the magnetoresistive approach to determine feasibility.

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H. SUPERCONDUCTIVE CONVERTERS

1. General Considerations

Superconductivity is a low temperature property common to several metals whereby a reversible transition from normal electrical resistance to zero resistance is brought about by variation of one or more of the following quantities:

(a) Temperature
(b) External magnetic field
(c) Internal electric current

When in the superconducting state the metal possesses, in addition to its zero resistance, the property of zero magnetic induction (exclusion of applied magnetic field). Both properties are expected to find extensive application in the future.

The application of superconductivity to d-c voltage conversion and amplification has not, within the author's knowledge, been considered previously. It would appear likely that the resistance transition is the important factor in this application. Special consideration is given here to the use of this resistance change.

2. Superconductive Switching

The superconductors can be switched by any of the above three factors or combinations of them. The cryotron is a superconductive resistance switch consisting of two elements. The resistance of one element, known as the gate, is controlled by application of current tothe second, the control element. The current in the control element furnishes the magnetic field necessary to cause the second element to become normally resistive. The action of a cryotron is illustrated schematically in Figure 33, in which I_C and I_G are respectively the control and gate currents. The cryotron may be either of the wire-wound type shown in Figure 33, or the crossed-film type in Figure 34. The latter is generally preferable from the standpoint of switching speed. Films may also, in principle, be deposited with small thicknesses to achieve higher resistance. The latter is an important consideration, since, as shown on Figure 28, the efficiency of a conversion circuit using resistance switching is directly dependent on the ratio $R_2:R_1$, where R_2 is the high and R_1 the low value of the switched resistance. At first sight, this would seem favorable to cryotron switching, as the resistance in the superconducting state is zero. However, both R_1 and R_2 must include the lead resistance as well as the internal resistance of the voltage source. If either of these is comparable to the normal gate resistance, the efficiency of cryotron switching is reduced. Hence lies the need for high gate resistance.

Switching from the superconductive state to the normal state might also be accomplished by other means. One scheme might utilize saturation of the output transformer as a switching mechanism. In this scheme, saturation of the output transformer (which may have a superconducting primary) will cause excessive primary currents exceeding the critical value for superconductivity. This would switch the superconducting element to its higher resistance state. Various types of feedback arrangements might also be used including a separate saturating transformer, magnetic fields, or thermal effects. The feasibility of these switching arrangements is not known and further work would be required to develop the circuitry and prove feasibility. The purpose of this initial investigation is to determine basic feasibility rather than develop circuitry or mechanisms. Therefore the particular switching arrangements will not be considered in great detail. Basic feasibility can be determined by examining and calculating the lumped parameter efficiencies of the specific problem areas. Thus the overall efficiency is:

$$\eta = \eta_T \cdot \eta_R \cdot \eta_{HRf} \cdot \eta_{SRRf} \cdot \eta_{CRf} \tag{21}$$

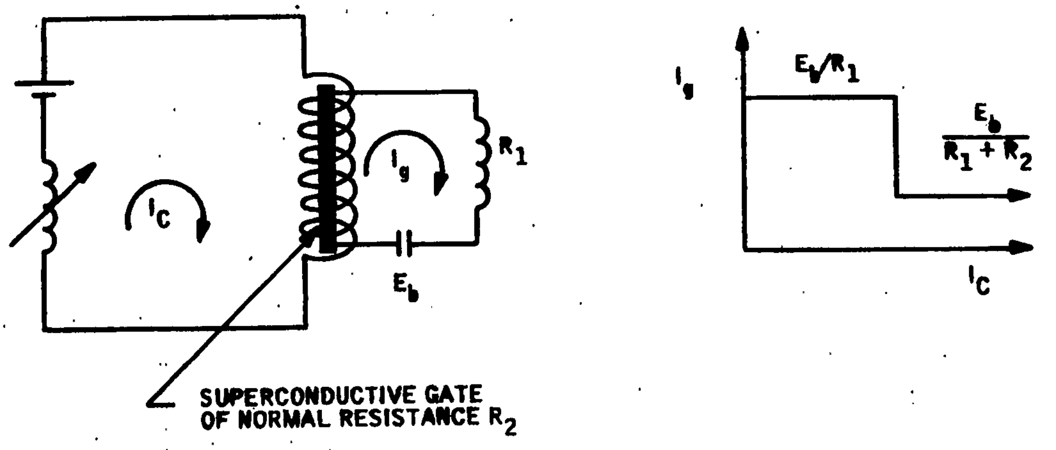

Figure 33 - CRYOTRON CHARACTERISTICS

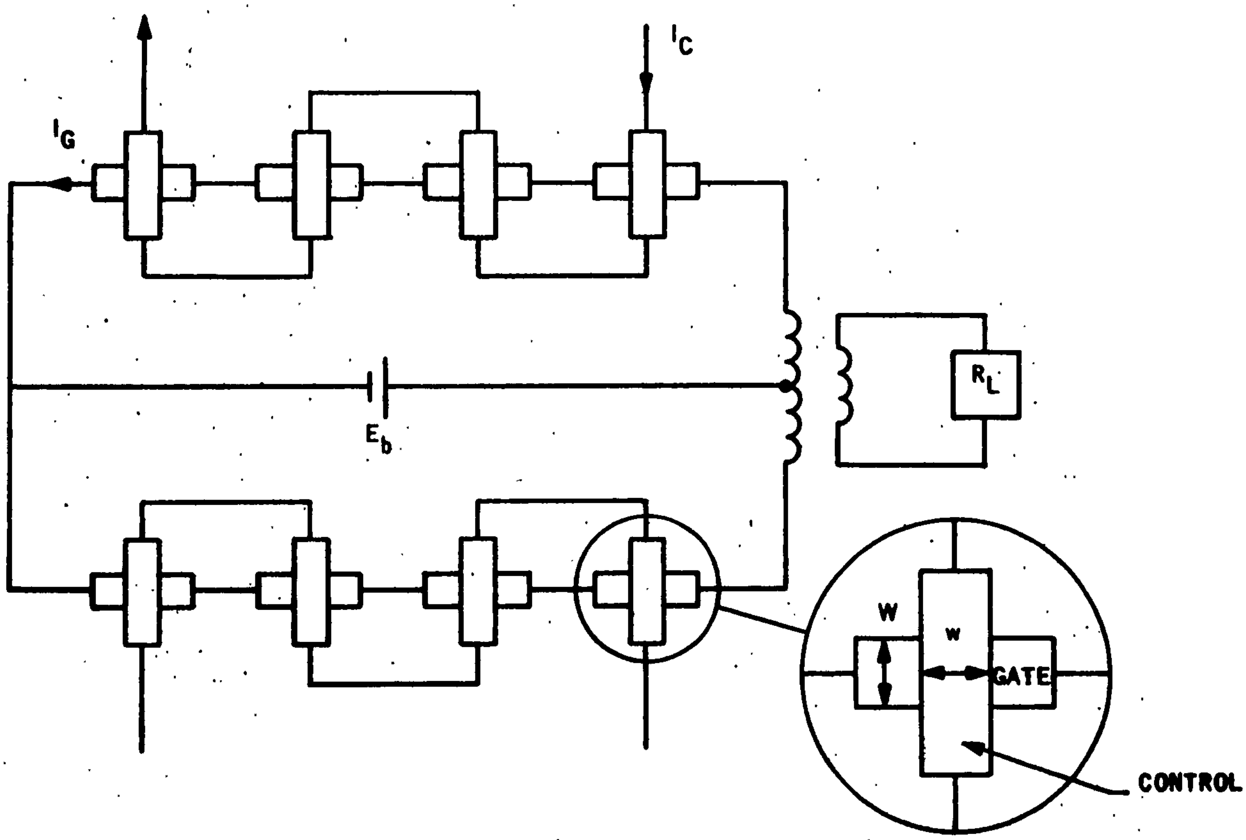

Figure 34 - APPLICATION OF CROSSED-FILM CRYOTRON TO VOLTAGE CONVERSION

where: η	=	Overall efficiency (75% desired)
η_T	=	Transformer efficiency (96% estimated)
η_R	=	Rectifier and filter efficiency (96% estimated)
η_{HRf}	=	Header lead heatflow and compensating refrigeration efficiency factor
η_{SRRf}	=	Internal transducer loss and compensating refrigeration efficiency factor
η_{CRf}	=	Thermal insulation efficiency of the superconductor container
		(Transducer switching and switching control losses are assumed to be small and neglected)

Two of the items η_T and η_R have been estimated to have relatively high efficiency and will not be considered further. η_{HRf}, and η_{SRRf} have considerable loss and must be considered in detail.

The heat flow into the cryotron through the header leads is a serious problem which reduces overall device efficiency. To minimize this problem optimum leads must be selected. The optimum copper lead pair has been shown by McFee to have a heat flow of 0.084 watt per ampere of current carried. * Calculations have been made to determine refrigeration power required over the input voltage range. These calculations are shown in Appendix D. Computations show that the refrigeration power required based upon the Carnot cycle is 68.1 watts per watt of heat flow into the cryostat (assuming the refrigerator efficiency is 100%). With the optimum leads this results in a refrigeration requirement of 5.72 watts per ampere carried through the cryotron. This assumes single stage optimum copper leads. It has been reported in the literature ** that the theoretical required refrigeration power

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* R. McFee "Optimum Input Leads for Cryogenic Apparatus" The Review of Scientific Instruments, Feb. 1959

** R. McFee - "Application of Superconductivity to the Generation and Distribution of Electric Power." Electrical Engineering Feb. 1962.

might be reduced to 1.0 watt per ampere carried if cascaded leads are used with multistage refrigeration. This is the most optimistic estimate found. This high refrigeration power requirement per ampere indicates that the lumped efficiency factor η_{HRf} for this particular loss will decline with lower source voltages. η_{HRf} has been calculated and plotted for various input voltages and shown in Figure 35. The greater current requirement at lower source voltages causes η_{HRf} to decline very rapidly. The maximum theoretical efficiency is only 50% at 11.6 volts and declines to zero at 5.72 volts for the single stage lead. This voltage is considerably above our 0.1 to 1.5 volt range and hence makes this approach appear unfavorable. If cascaded leads with multistage refrigeration were used, the maximum theoretical η_{HRf} efficiency factor would be 60% for a 2.5 volt source and 35% for a 1.5 volt source. This is still unfavorable. Since these curves assume 100% efficiency for the refrigerator and since this η_{HRf} factor must be multiplied by the other efficiency factors, this parameter alone rules out this approach for low voltage energy conversion.

Calculations have been made to determine the effect of the transducer losses and to estimate the (R2:R1) resistance ratio required between the normal and the superconducting element considering optimum lead and source resistance effects. These calculations shown in Appendix D include refrigeration and assume that refrigeration power will be obtained from the converted device output at 100% efficiency. The calculations show that the transducer must be 98.6% efficient to achieve any output power whatsoever. This requires a transducer resistance ratio of over 10,000 to 1 to obtain an output. For any appreciable efficiency the ratio must be much greater. Since this must include the lead resistance these ratios are very difficult to obtain.

Calculations have also been made on the assumption that refrigeration could be supplied separately in the form of liquid helium. The liquid helium would have to be carried with the device and would contribute to weight and volume similar to fuel requirements. These calculations show that the weight of

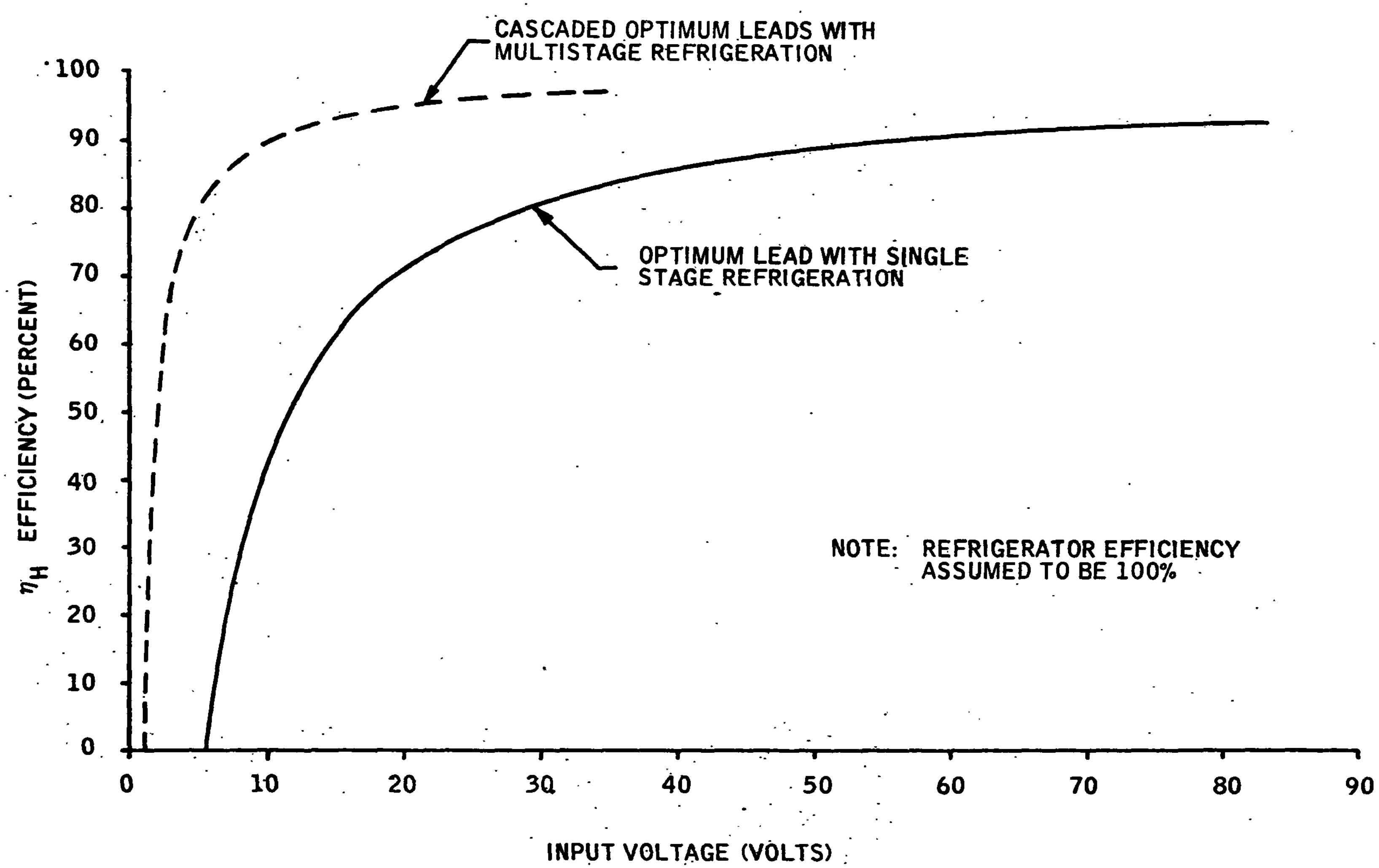

Figure 35 - THEORETICAL MAXIMUM EFFICIENCY FACTOR η_{HRF} DUE TO HEAT FLOW AND COMPENSATION, REFRIGERATION REQUIRED FOR OPTIMUM CRYOTRON INPUT LEADS VS. INPUT VOLTAGE

helium for 8 hours of operation, considering only one of several losses, is 11.3 pounds. Since other losses and a container is involved, the required weight and volume make this approach impractical for a portable device.

Calculations shown in Appendix D indicate that the required dimensions of a superconducting material may be reasonable for extremely low input voltages. For the higher input voltages the physical dimensions become more difficult to obtain or more complex. The large number of series connected cryotrons at the higher voltages would reduce reliability. Other factors mentioned above preclude the use of low input voltages. The higher input voltages are not of interest in this program.

3. Conclusions

The several factors mentioned above indicate that the superconducting approach as such does not appear feasible for low input voltage conversion with the present state of the art. There does not appear to be any immediate means of improving this situation because the low input voltage requires extremely heavy input leads resulting in high thermal losses which limit overall efficiency.

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I. PHOTOCONDUCTIVE APPROACH

The basic characteristics of this scheme are to employ photoconductive switches in a circuit in such a way as to modulate the d.c. input. The modulated voltage would be stepped up by a transformer, and the higher voltage subsequently rectified. Besides the switching circuit, a source of modulated light must be available. One possible circuit is shown in Figure 36.

In order that the efficiency of the system should be of the order of 90%, Figure 28 shows that the ratio of dark to light resistance of the photoconductive elements should be approximately 1000. Of all the known photoconductive materials useful at room temperature, only CdS and CdSe exhibit change of resistance of this order. Of the two, CdS is more photosensitive, but has a slightly slower speed of response than CdSe. The following discussion is limited to CdS but could be applied with only a few changes to CdSe.

1. Activation of Cadmium Sulfide

Although pure CdS is a rather insensitive photoconductor, it can be activated by the addition of certain impurities to make it extremely sensitive. Changes of more than five orders of magnitude between dark and illuminated resistivity are readily obtained. In this section we will give a brief discussion of the influence of certain impurities on the photoresponse.

Cadmium sulfide is a Group IIB-Group VIA semiconductor having a bond characterized by a high degree of ionicity. Thus the cadmium and sulfur atoms are held together in the lattice by valence bonds formed by the transfer of the two valence electrons in a cadmium atom to an adjacent sulfur atom. Sulfur, having six valence electrons, adds the two transferred electrons to its outermost shell, completing it (see Figure 37).

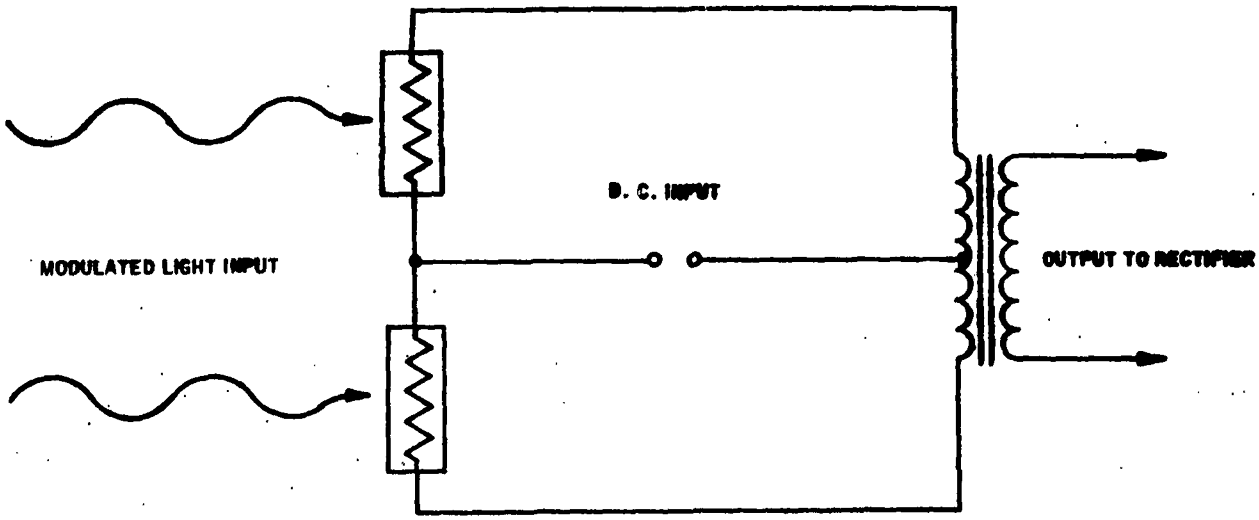

Figure 36 - BASIC PHOTOCONDUCTIVE SWITCHING CIRCUIT

Cd^{2+}	S^{2-}	Cd^{2+}	S^{2-}
S^{2-}	Cd^{2+}	S^{2-}	Cd^{2+}
Cd^{2+}	S^{2-}	Cd^{2+}	S^{2-}
S^{2-}	Cd^{2+}	S^{2-}	Cd^{2+}
Cd^{2+}	S^{2-}	Cd^{2+}	S^{2-}
S^{2-}	Cd^{2+}	S^{2-}	Cd^{2+}

Figure 37 - REPRESENTATION OF CdS LATTICE

The energy necessary to free one of these bound electrons is about 2.4 electron volts. If a photon of energy equal to or greater than 2.4 electron volts, (i.e., wave length less than or equal to $\lambda = \frac{1.24}{2.4} = 0.5\mu$), is absorbed by the lattice, it will free an electron and hole pair. Since the electron lifetime is about 10^{-3} seconds, whereas the hole lifetime is less than 10^{-9} seconds, only the electrons contribute appreciably to the increased conductivity. Figure 38 illustrates the energy level diagram of pure (intrinsic) CdS.

Activation of CdS by copper and chlorine, by copper and gallium, or by copper and indium increases the photosensitivity thousands of times over that of pure CdS. Consider what happens when chlorine is added as an impurity to the lattice. Chlorine enters substitutionally at a sulfur site. Since it has seven valence electrons, the chlorine can accept only one electron from a neighboring cadmium atom. The second electron is then very slightly bound to the lattice site, with a binding energy of about 0.01 ev. This is sufficiently low so that the site will be ionized at room temperature. Thus CdS incorporating chlorine substitutionally has a large electrical conductivity due to the free electrons donated by the chlorine-cadmium centers. Copper, on the other hand, having one electron in its outermost shell, enters substitutionally at a cadmium site. Since copper can give up only one electron, the adjacent sulfur site will have only seven electrons. This site will then trap any free electrons which wander to its vicinity. This reduction of the free electron concentration will therefore reduce the electrical conductivity. The lattice incorporating copper and chlorine is shown in Figure 39.

If the copper and chlorine are added in approximately equal amounts, the chlorine sites will donate free electrons to the lattice, which will then be trapped by the copper sites. These trapped electrons can be freed by a photon having an energy of approximately 2.0 ev (0.62μ wave length) or more. The energy level scheme is shown in Figure 40.

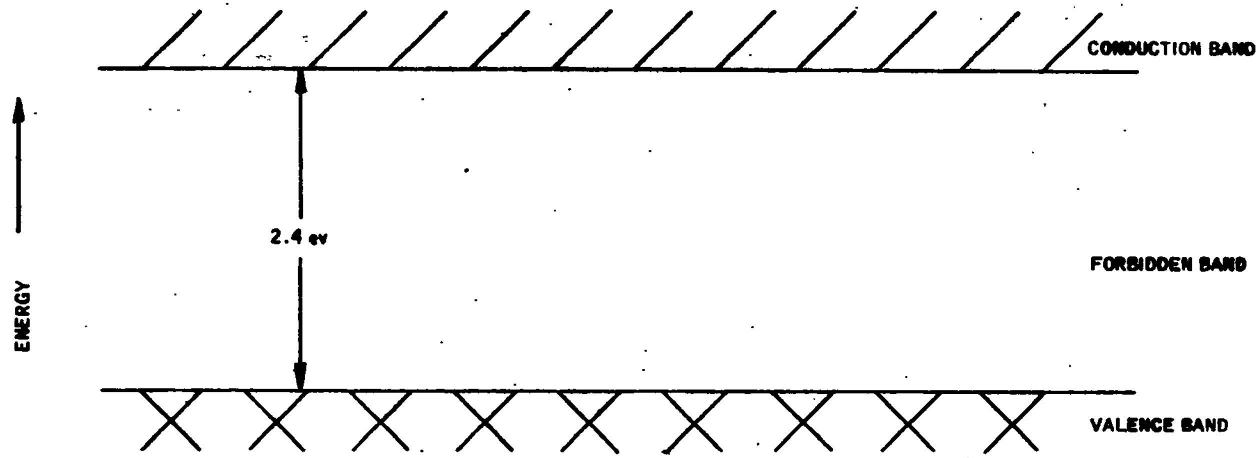

Figure 38 - ENERGY LEVELS IN PURE CdS

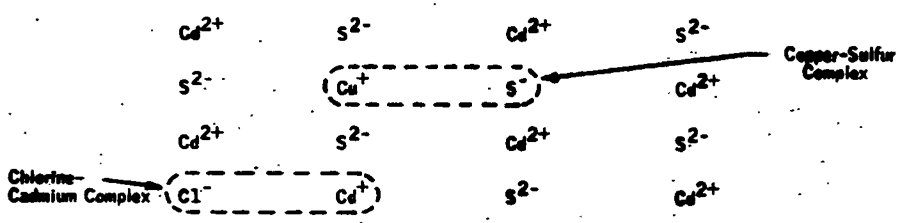

Figure 39 - COPPER AND CHLORINE ACTIVATED CADMIUM SULFIDE

It can be shown* from arguments based on conservation of electrical charge and conservation of lattice sites that charged and uncharged lattice vacancies must also exist. They shall be ignored here since impurity activation can be discussed at this level without referring to them.

For the other common activating agents in CdS, indium and gallium, arguments similar to those above may be applied. Both In and Ga substitute for cadmium of lattice sites, donating free electrons in the same manner as chlorine. Thus CdS may be activated with approximately equal amounts of copper and indium or copper and gallium. Optimum concentrations are of the order of several hundred parts per million.

The spectral responses of unactivated and activated CdS are shown in Figure 41.

2. Phenomenological Description of Photoconductivity in Cadmium Sulfide

Cadmium sulfide is one of a group of semiconductors having wide energy gaps known as "photoconducting insulators". In this section, we shall describe briefly some of the important features of the photoconductive process in such materials.

A basic expression describing photoconductivity relates the equilibrium number of free carriers N to the generation rate g and lifetime τ.

$$n = g\tau. \tag{22}$$

Equation (22) applies not only to photoconductors but to any generation-recombination process. For example, the equilibrium population N of a nation is given by the product of the birth rate g and the life expectancy τ.

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* Kroger, Vink, and van den Boomgaard, Z. physik, Chem. 203, 1 (1954).

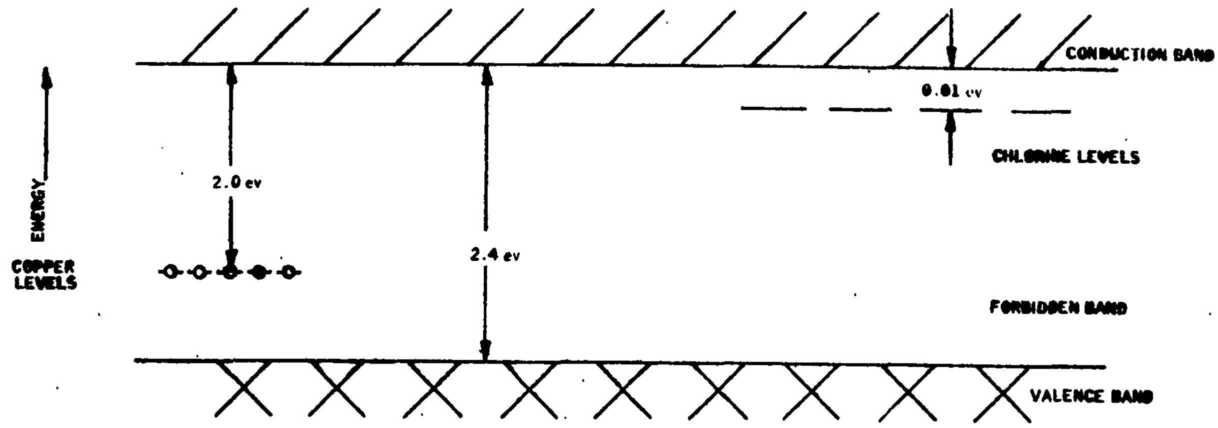

Figure 40 - ENERGY LEVELS IN Cu, Cl ACTIVATED CdS

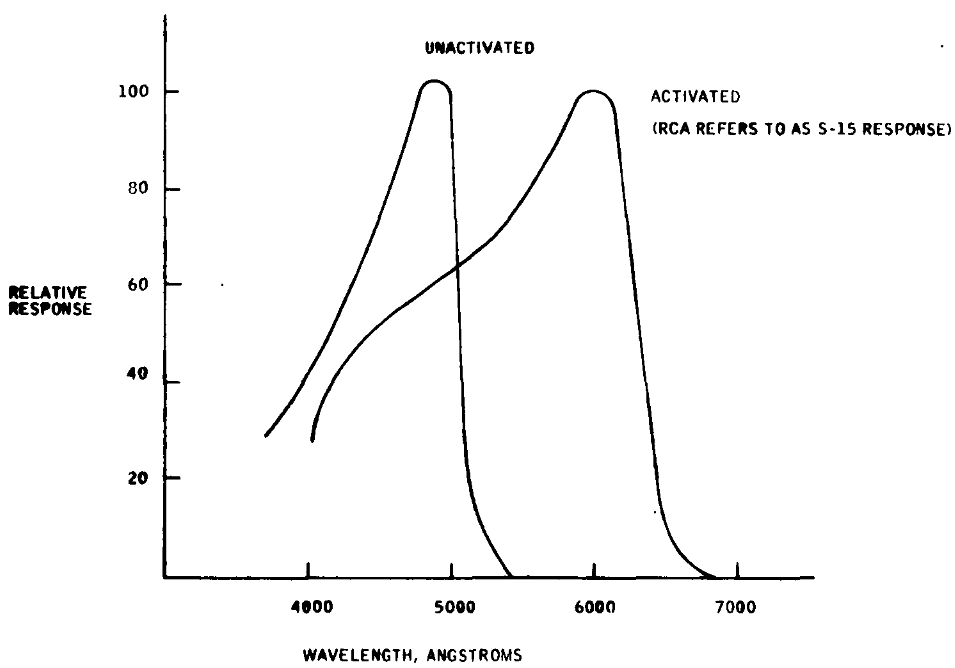

Figure 41 - RELATIVE RESPONSES OF ACTIVATED AND UNACTIVATED CdS

Consider the motion of the free carriers through a crystal. They drift in an applied field (diffusion will be neglected) at a velocity v determined by the product of their mobility μ and the field strength E.

$$v = \mu E = \mu \frac{V}{d} , \tag{22}$$

where V is the applied voltage and d the distance between electrodes. The current density J is given by

$$J = nev , \tag{23}$$

where e is the electronic charge and n the carrier density. The total current I through the sample is related to J by

$$I = Jwh , \tag{24}$$

where w is the width and h the thickness of the semiconductor. Assuming that each incident photon liberates one electron (unit quantum efficiency) and that the hole lifetime is negligible compared to the electron lifetime valid for CdS, we obtain by combining Eqs. (21), (22), (23), and (24)

$$I = \frac{ne\mu Vwh}{d} = \frac{N}{d} e\mu\frac{V}{d} = g\tau\, e\mu \frac{V}{d^2} . \tag{25}$$

The carrier transit time t between electrodes is

$$t = \frac{d}{v} = \frac{d^2}{\mu V} . \tag{26}$$

Thus the current is given by

$$I = eg\ \tau/t \tag{27}$$

Examining Eq. (27) we note that the current is not given simply by the product of the number of electrons released per second g with the electronic charge, e, but also involves the ratio τ/t. If τ/t is much larger than unity, that is, the carrier lifetime is much greater than the transit time between electrodes, then a large photoconductive current can flow. As the electron freed by the photon reaches the positive electrode, leaving the photoconductor, an electron enters at the negative electrode. When this electron reaches the positive electrode, another will enter at the negative electrode, and so on for τ/t transits. Thus if τ/t is, say, 1000, then 1000 electrons will flow through the external circuit for every one freed by a photon. This amplification factor is responsible for the large photoconductive response in CdS. Since it is directly proportional to the carrier lifetime, it is apparent that high photosensitivity is accompanied by long lifetime, i.e., poor frequency response. Note that this is only possible if electrons can enter and leave the electrodes without difficulty. If the contacts are nonohmic (rectifying), then potential barriers at the electrodes will not permit electrons to easily enter the photoconductor and will therefore reduce the gain.

3. Performance of Commercially Available CdS Photocells

The following table and graphs give some of the operating characteristics of an RCA type 4424 photoconductive CdS cell. This is a relatively new model and the specifications are quite typical of other makes.

TABLE III

OPERATING CHARACTERISTICS OF AN RCA
TYPE 4424 PHOTOCONDUCTIVE CdS CELL

General:		
Spectral Response		S-15
Wave length of Maximum Response		5800 ± 500 angstroms
Sensitive Surface, including metallic Electrodes:		
Shape		Rectangular
Length (Minimum)		0.22 in.
Width (Minimum)		0.22 in.
Area (Minimum)		0.048 sq. in.
Maximum Length (Excluding flexible leads)		1.2 in.
Greatest Diameter		0.656 in.
Maximum Axial Distance from Flat End of Envelope to Sensitive Surface		0.08 in.
Operating Position		Any
Weight (Approx.)		.13 oz.
Minimum Ratings, Absolute-Maximum Values:		
VOLTAGE BETWEEN TERMINALS (DC or PEAK AC)	110 max.	volts
POWER DISSIPATION -		
Sensitive surface fully illuminated: Continuous service	0.2 max.	watt
Sensitive surface partially illuminated: Continuous service	4.2 max.	watt/sq.in.
PHOTOCURRENT	50 max.	ma
AMBIENT TEMPERATURE RANGE	-75 to + 60	^{o}C

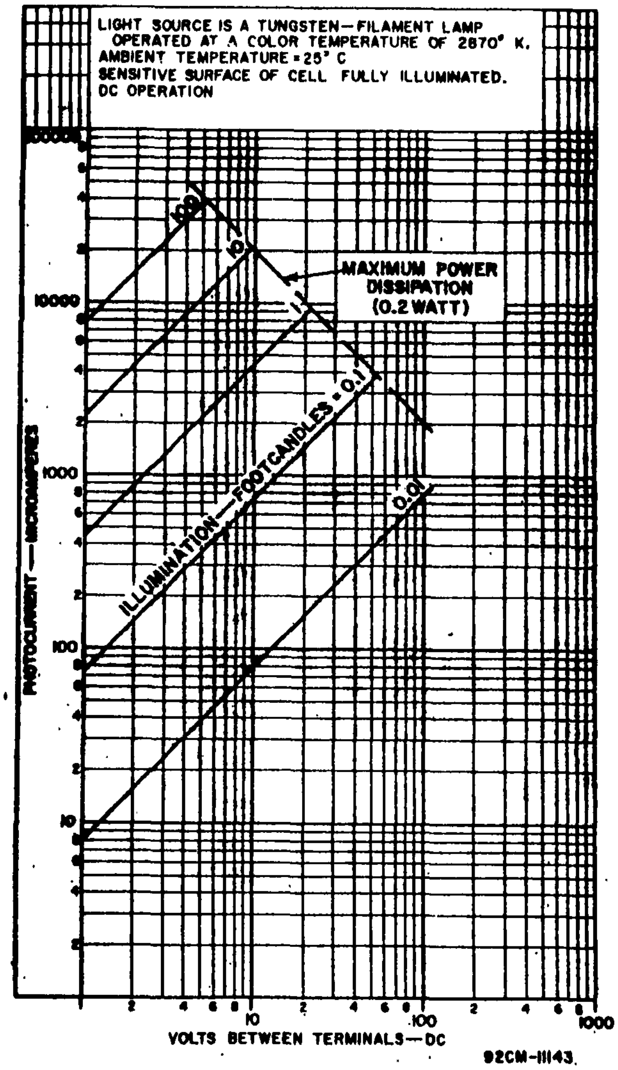

Figure 42 - AVERAGE CHARACTERISTICS OF TYPE 4424

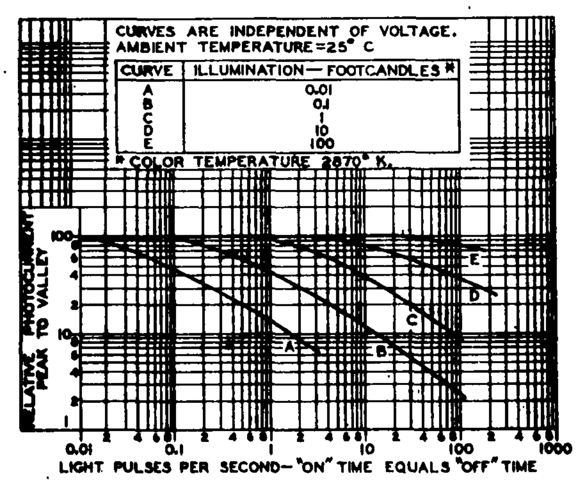

Figure 43 - RESPONSE CHARACTERISTICS OF CADMIUM SULFIDE CELL TO PULSED LIGHT

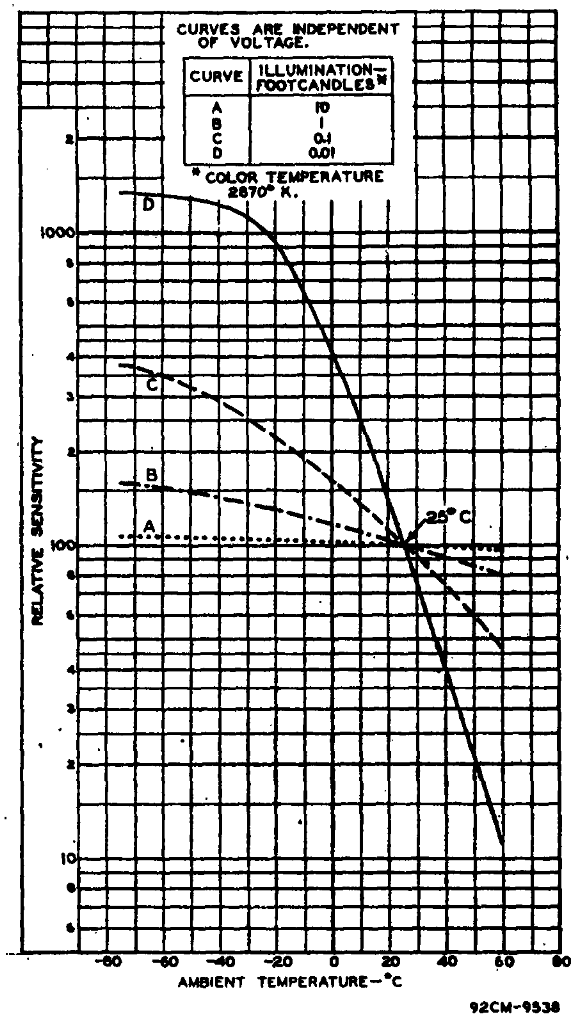

Figure 44 - TYPICAL CHARACTERISTICS OF CADMIUM SULFIDE CELL

From this data, we can compute some of the characteristics of our photocell. Suppose that the d-c input is 1.0 volt at 100 amps. With a bias of one volt and illumination of 100 ft. c., a 4424 will handle a photocurrent of 8 ma (Figure 42). Therefore, about 12,500 cells would be required in order to handle 100 amps. Since the sensitive area of each cell is 0.048 in^2, we have a sensitive area of 600 in^2 or approximately 4.2 ft^2. The resistance of this parallel array would be 0.23 ohm when illuminated and 710,000 ohms in the dark. This "on" resistance is high for our application.

Figure 43 shows the frequency response of the cell. At 100 foot candles illumination, a dark to light resistance ratio of 100 is possible at about 20 cps. In order to achieve a ratio of 1000, and therefore 90% efficiency, much slower rates would have to be employed. Increasing the light level would improve this situation somewhat.

Information on an RCA developmental type CdSe cell (No. C7218) indicates an area of about 58 ft^2 and a resistance of 0.01 ohm would meet the requirements for current carrying capacity. In addition, the response is about five times as fast.

4. Conclusions

The outlook for this method would not appear to be too favorable. While some reduction in area should be possible for CdS, and perhaps quite a large reduction for CdSe, it appears that the slow response of these materials would be a severe handicap in achieving the desired resistance change at a reasonable frequency. Spacing the electrodes closer to reduce the resistance and the necessary area would also shorten the response time slightly, but not enough to alter the characteristics appreciably.

Experience at Honeywell Brown Instrument Division, using a Claire 603 AL photoconductor in this circuit, excited by a neon glow lamp pulsed at 60 cps, indicates an effective resistance ratio of between 10:1 and 100:1. The ratio would be over 1000:1 at frequencies of the order of 0.1 cps.

It might be possible to increase the speed of response of the photoconductor by control of the doping; however, this would be at the expense of reduced photosensitivity. Whether an overall improvement in performance would be attained is problematical. Another method of achieving a faster response is due to Borkan and Weimer*. In this method, two photocells are used, one with a smaller and slower response than the other. When the outputs of these two cells are subtracted, an improvement is found in both the rise and decay response time, although the total response is smaller.

The possibility of using other materials should not be overlooked, although no other photoconductors presently being used will give nearly the desired change in resistance. One material which should be investigated is $CdIn_2S_3$. This material was reported** to have a low resistivity and large photosensitivity, as well as a time constant shorter than CdS. Later information‡, however, indicates that $CdIn_2S_3$ has not lived up to its earlier promise.

In general, then, it would seem that photoconductive switching with CdS or CdSe is suitable for systems where a basic efficiency of about 50% is permissible, and where the large area of the photoconductor is not a handicap. The problem of supplying a modulated high intensity light source has not been discussed, since this is so dependent upon the application.

Photoresistive Section prepared by James Garfunkel

James Garfunkel
Research Scientist

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* Borkan, H. and P. K. Weimer, RCA Rev. XIX:1 62-76 (March, 1958).
** Kodmans, H. and H. G. Grimmeiss, Physica 25, 1287 (1959).
‡ Private communication.

SECTION V

CONCLUSIONS

Examination of the conclusions of each preceding section shows that the tunnel diode approach, Hall effect approach, superconductive approach, and the photoresistive approach do not appear promising for low input voltage high efficiency power conversion. The electromechanical approach and the magento-resistive approach will require further effort to determine feasibility. The transistor approach appears feasible if a sufficient number of transistors is used or if superior low saturation high current transistors are developed. Investigation of the liquid metal magnetohydrodynamic approach will commence in the next quarter. New approaches should be investigated if they show any promise. Transistor parameter measurements have been initiated and experimental verification of the transistor approach should also be initiated. This should be directed toward a converter incorporating both voltage and current drive.

During the next quarter, effort should be directed toward determining the optimum input voltage ranges for the more promising approaches. For example, the transistor approach is quite limited for inputs of less than 0. 7 volt.

Required transducer parameters should be calculated for operation from higher voltages obtained by connecting a few source cells in series.

A. PROGRAM FOR THE NEXT INTERVAL

During the next interval, investigation of present favorable approaches will continue and effort will be directed toward finding new approaches. The review of literature being received as a result of our literature search will continue and all applicable useful information on transducer materials and devices, circuitry, and approaches will be retained. Investigation of the liquid metal magnetohydrodynamic approach will commence. Any new approaches which appear to have significant possibilities will be investigated.

Evaluation of selected high current transistors will continue and be completed shortly. Effort will be directed toward keeping abreast of the high current transistor state of the art and procuring the best available transistors for this application. An experimental transistor power amplifier breadboard will be constructed with these transistors to evaluate transistor drive requirements and switching losses under the actual conditions of operation. This breadboard will be operated with source voltages between 1. 0 and 1. 5 volts. Combinations of current and voltage drive will be used. Performance measurements with various source impedances will be made to determine the effects of source impedance on switching characteristics.

Effort on the tunnel diode approach will be continued at a low rate of effort. Since this approach appears sub-marginal, if the desired efficiencies are to be obtained, effort on this approach will be directed at a more accurate determination of the state of the transducer art and keeping abreast of latest developments. If the desired efficiency requirements are reduced or if new state of the art developments occur, the rate of effort on this approach might be increased.

Effort on the electromechanical approach will continue. This effort will be directed toward the evaluation of configurations which are not limited by gravitational fields. Several configurations will probably be investigated.

Further calculations will be made on the magnetoresistive approach. This approach appears marginal and further effort should result in a more accurate determination of feasibility.

The Hall effect approach, the superconductive approach, and the photoresistive approach do not appear favorable and hence no appreciable effort is planned on these approaches during the next quarter. Interest will be retained in these approaches and should new developments occur, the feasibility of these approaches will be re-evaluated.

B. IDENTIFICATION OF KEY TECHNICAL PERSONNEL

Resumes of the personnel assigned to this program are as follows:

W. L. HUNTINGTON, Chief Engineer, Product Development

Experience

- Presently Chief Engineer for Product Development working on development of missile safety, arming and fuzing systems, adaption kits and components, fuzing system and adaption kit training devices, turret and fire control systems, munitions, cryptographic devices, power supplies, thermal batteries, hydraulic components and transistor devices.

- At Aeronautical Division, he served as a design engineer on an automatic pilot and on automatic temperature control and flight engine control components, and as a project engineer in the development of air ram switches, altitude warning switches, differential pressure switches, and servo motors.

Professional Background

- Stout Institute, Wisconsin
- University of Minnesota
- BS, Architectural Engineering, Iowa State College

L. E. ALBERTS, Project Supervisor

Experience

- Presently design engineering supervisor of Ordnance Engineering Equipment Section specializing in the design and development of transistor power supplies, inverters, converters and associated solid-state electronic devices.

- Twenty-seven years experience with the Honeywell Corporation, particularly in electric assemblies, as a design engineer, a field engineer, and as a supervisor.

Professional Background

- Registered Electrical Engineer, State of Minnesota

J. T. LINGLE, Project Engineer

Experience

- Specialized in design and development of solid-state power supplies.
- Experience dating back to 1952 on applications of transistors, power converters, switching circuits, and voltage regulators.
- Active experience on such projects as a 1200-watt thermoelectric generator and a transistor switching study for the U. S. Signal Corps.

Professional Background

- BSEE, University of Minnesota
- Registered Professional Engineer, State of Minnesota

G. D. LONG, Section Head, Solid State Physics

Experience

- Honeywell Research Center. Worked on:
  - electrical properties of semiconductors.
  - use of electrical measurements to evaluate purity and perfection of crystals of semiconductor materials.
  - solid state devices and electronic components.

- theory of electrical properties of p-n junctions.
- electrical measurements and thermometry at very low temperatures (liquid helium range).
- diffusion of impurities into semiconductors.
- properties of Hall effect and magnetoresistance type devices.
- studies of electron scattering in semiconducting materials, properties of semiconductor strain gauges and related devices.
- research and development in advanced microelectronics.

Professional Background

- BS, Physics, Lehigh University
- PhD, Physics, University of Pennsylvania, Thesis: "Studies of the Effects of Pressure on the Electrical Properties of Semiconductors"

Publications

"Stress Dependence of the Piezoresistance Effect, " J. Appl, Phys, 32 2050 (1961)

"Scattering Anisotropies in n-Type Silicon" (with J. Myers), Phys, Rev. 120, 39 (1960).

"Scattering of Conduction Electrons by Lattice Vibrations in Silicon," Phys. Rev. 120, 2024 (1960).

"Hall Effect and Impurity Levels in Phosphorus-Doped Silicon" (with J. Myers) Phys. Rev. 115 , 1119 (1959).

"Ionized-Impurity Scattering Mobility of Electrons in Silicon" (with J. Myers) Phys. Rev. 115, 1107 (1959).

"Impurity Compensation and Magnetoresistance in p-Type Silicon" (with C. Motchenbacher and J. Myers), J. Appl. Phys. 30, 353 (1959).

R. D. FENITY, Section Head, Ceramics

Experience

- Present work on oxide thermoelectrics and thermistor materials and single crystal studies of oxide semiconductors.
- Engaged in development and research on dielectric and piezoelectric titanates and niobates.
- Investigation of ceramic insulator bodies with special properties for improved ceramic-to-metal seal production and oxide semiconductors for thermistors, thermoelectric generators, heaters and other applications.

Professional Background

- BS, Ceramic Engineering, University of Illinois

Publications

"Possible Explanation of Positive Temperature Coefficient in Resistivity of Semiconducting Ferroelectrics", Journal of the American Ceramic Society, 44, 249 (1960).

"Oxide Thermoelectric Generators", Electronics, February 2, 1962.

O. S. LUTES, Principal Research Scientist

Experience

- Presently working on low temperature thermometry and superconductive refrigeration.
- Previous work in low temperature physics:
    - electrical, magnetic and thermal properties of metals and alloys at temperatures of liquid helium.

- magnetic transitions in dilute alloys.
- liquid helium (isotope 3) refrigeration for temperatures below 1° K.
- techniques for ultra-sensitive magnetization measurements.
- effective internal magnetic fields in alloys.
- ultra-purification of metals.
- boundary scattering phenomena in pure aluminum.

- With National Bureau of Standards, studies of magnetic properties of superconductors in the liquid helium temperature range. Collaborated in investigation of thermal effects in solid materials condensed at 4.2° K from high frequency nitrogen discharge.

Professional Background

- BS, Physics, Carnegie Tech
- MA, Physics, Columbia University
- PhD, Physics, University of Maryland. Thesis: "Superconductivity of Microscopic Tin Filaments".

Publications

"Galvanometer Deflection Micrometer", Rev. Sci. Instr. 31, 780 (1960).

"Superconductivity of Microscopic Tin Filaments", Phys. Rev. 105, 1451 (1957).

"Abundance of Free Atoms in Solid Nitrogen Condensed at 4.2° K from a Gas Discharge", J. Chem. Phys. 24, 484 (1956).

"Superconducting Transitions in Tin Whiskers", Phys. Rev. 97, 1718 (1955).

O. N. TUFTE, Senior Research Scientist

Experience

- Honeywell Research Center, Solid State Physics Group. Worked on:
  - diffusion of impurities in semiconductors.
  - silicon diffused transistors and other diffused devices.
  - transistor choppers.
  - epitaxial growth of silicon and deposition of silicon on inert substrates.
  - piezoresistance effect in semiconductors and its application to semi-conductor strain gages and pressure sensing devices.
  - studies of energy band structures of semiconductor materials using galvanomagnetic measurements.

Professional Background

- BA, Physics, Chemistry, St. Olaf College
- PhD, Physics, Northwestern University. Thesis: "Growth and Semiconducting Properties of Gray Tin Single Crystals".

Publications

"Magnetoresistance of Oriented Gray Tin Single Crystals", (with A. W. Ewald) Phys. Rev. 122, 1431, (1961).

C. D. MOTCHENBACHER, Senior Research Scientist

Experience

- Electrical engineering consultant to the corporate Research Center and as such has extensive experience in circuit design for a wide range of applications:

- Special experience in working with new type sensors and experimental equipment.

  - special background in low-level, low-noise amplification systems

- Two years in the field of cryogenics on experiments with the Hall effect and other transport phenomenon in semiconductors.

- Research in the development of industrial and residential temperature controls.

Professional Background

- BSEE, South Dakota State College
- Graduate courses at University of Minnesota

Publications

"Impurity Compensation and Magnetoresistance in P-Type Silicon". D. Long, C. D. Motchenbacher and J. Meyers, Journal of Applied Physics, Volume 30, March 1959.

"A Sensitive Displacement Meter Utilizing a Hall Effect Probe", C. D. Motchenbacher and S. B. Schuldt - AIEE, DP61-591 (1961).

"A Professional Guidance Program for High School Science Students", C. D. Motchenbacher - AIEE, DP61-590 (1961).

J. T. MAUPIN, Senior Research Engineer

Experience

- Honeywell Research Center, doing research in microelectronic theory.
- Honeywell Semiconductor Division; worked in design, development and application of semiconductor devices, mostly power transistors.
- Three years as supervisor, the latest being manager of Applications Engineering.
- Four U. S. Patents and several publications, mostly in the field of solid state devices and circuits.
- Lecturer in Electrical Engineering at University of Minnesota Extension School.

Professional Background

- BSEE, University of Kentucky
- MSEE, University of Minnesota
- Member of AIEE, Tau Beta Pi, Eta Kappa Nu

Publications

"The Interaxial Spacing and Dielectric Constant of Cable Pairs", Bell System Technical Journal, July 1951.

"Measuring Relative Phase Shift at VHF:, Bell Labs. Record, August 1955.

"The Tetrode Power Transistor", Transactions of the IRE Professional Group on Electron Devices, Jan. 1957.

"A Direct Coupled Linear Power Amplifier", Paper presented at 1958 Solid State Circuits Conference, Univ. of Penn., Feb. 20, 1958. Digest published in Conference Proceedings.

"Constant Resistance Transistor Stages", Transactions of IRE Professional Group on Circuit Theory, Dec. 1961.

J. E. ANDERSON, Staff Physicist, Project Scientist

Experience

- Presently Staff Physicist with the Honeywell Military Products Group Research Department. Previous work included:

  - two and one-half year leave of absence from Honeywell to take Ph. D. at M. I. T. Course work was primarily in the areas of special and general relativity, atomic and nuclear physics, tensor analysis, physics of high speed flow, magnetohydrodynamics, upper atmosphere physics, plasma physics, advanced applied mathematics, quantum theory of matter, electromagnetic theory, statistical theory of gases and fluid mechanics.

  - with Honeywell, key analyst in dynamic flight controls and inertial navigation systems.

  - with NACA, engaged in theoretical analysis of structures including problems in stress, deflection and vibration of solid plates by means of variational techniques and problems in aeroelasticity.

Professional Background

- BSME, Iowa State University, 1949
- MSME, University of Minnesota, 1955
- PhD, Massachusetts Institute of Technology, 1962
- Thesis subject: Stability and Structure of Magnetohydrodynamic Shock Waves.

J. H. GARFUNKEL, Research Scientist

Experience

- Honeywell Research Center, Primary Sensors Group since 1957.
  - worked on development of low temperature bolometers, improvement of infrared detector housings, development of photosensitive analog transducer and long wave length infrared detector.
  - developed specialized sensing systems for use in gas and vapor detection and an automatic inspection device for fuzes.
- High school and junior college physics and mathematics instructor.

Professional Background

- BS, Physics, St. John's University
- Graduate work, Montana State University and the University of Minnesota

DONALD E. BENZ, Associate Research Scientist

Experience

- Electrical Engineering Consultant at Honeywell Research Center since 1959
  - special experience in design of electronic equipment such as pulse generators and amplifiers for research experiments.
  - worked on electronic equipment utilizing the IR modulator developed at the Research Center.
  - experience on the low noise amplification project.

Professional Background

- BSEE, University of Minnesota
- Post graduate courses at the University of Minnesota

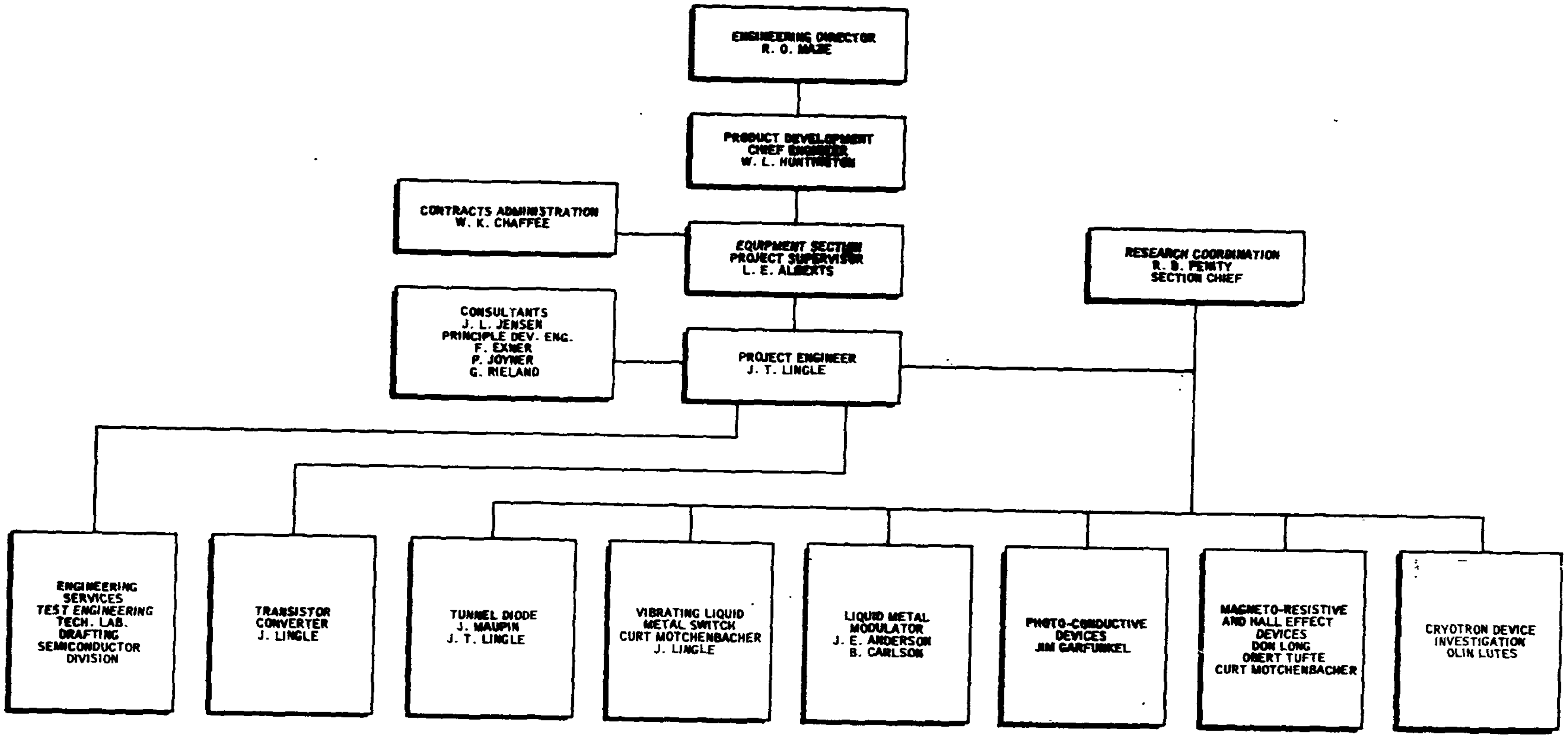

Figure 45 - PROJECT ORGANIZATION CHART

C. ENGINEERING TIME ANALYSIS

The engineering time devoted to this program during the first quarter (1 July to 30 September 1962) is broken down as follows:

John T. Lingle	Project Engineer	431.0 hours
G. D. Long	Section Head, Research Center	17.0 hours
R. D. Fenity	Section Head, Research Center	4.0 hours
O. S. Lutes	Principal Research Scientist	30.0 hours
O. N. Tufte	Senior Research Scientist	7.0 hours
C. D. Motchenbacher	Senior Research Scientist	31.5 hours
J. T. Maupin	Senior Research Engineer	20.0 hours
J. E. Anderson	Staff Physicist	31.0 hours
J. Garfunkel	Research Scientist	11.0 hours
D. E. Benz	Associate Research Scientist	33.0 hours
Evaluation Engineer		1.0 hour
Evaluation Tester		23.0 hours
Drafting		14.0 hours
Mechanical and Test Technician		9.0 hours
TOTAL ENGINEERING TIME		662.5 hours

APPENDIX A

CALCULATION OF TRANSISTOR PARAMETERS NECESSARY FOR FABRICATION OF LOW INPUT VOLTAGE CONVERTERS HAVING 75% EFFICIENCY

A. CALCULATIONS TO DETERMINE TRANSISTOR PARAMETERS REQUIRED FOR OPERATING A PUSH-PULL CONVERTER FROM A 1. 5 VOLT SOURCE

Since it is known that the transistor converters will perform better at higher voltages, the 1.5 volt upper limit will be considered first. The following assumptions will be made to determine the transistor parameters necessary to construct a device having 75% efficiency, ((η) = .75).

1. Assume transformer efficiency of 94% = η_T
2. Assume rectifier and filter efficiency of 96% = η_R
3. Assume an operating frequency of 1 kc = f
4. Output = 50 watts = P out
5. Transistor gain = 40 = h_{FE}
6. Input voltage, E_b, = 1.5 volts.

Determine required input current, (I):

$$I = \frac{P\,out.}{\eta Eb} = \frac{50}{.75 \times 1.5} = \frac{66.7 \text{ watts}}{1.5V} = 44.44 \text{ amperes} \tag{28}$$

Required efficiency of transistor and feedback circuitry.

$$.75 = \eta_o \times .94 \times .98 = (\eta_o)(\eta_T)(\eta_R) \quad \text{or} \tag{29}$$

$$\eta_o = \frac{.75}{.94 \times .96} = 83.2\%$$

Estimate the rise time at 20 μ sec. Estimate the fall time at 20 μ sec. From these estimates, determine drive power required and switching losses. Then estimate saturation voltage drop. For switching loss calculations the switching characteristics will be assumed to be as shown in Figure 46. In this figure, the switching "on" time = 20 μ sec. For switching "off" it is assumed that the current will remain constant for 10 μ sec. while the emitter to collector voltage (Vec) increases uniformly to 1.5 volts. It is then assumed that the current will decrease uniformly to zero while the voltage increases uniformly from 1.5 volts to 3.0 volts during the next 10 μ sec.

The switching losses can be obtained from equations (30), (31), and (32)* below:

$$P_1 = ft_{s_1} \left[1/6 \, (V_a I_b + V_b I_a) + 1/3 \, (V_a I_a + V_b I_b) \right] \tag{30}$$

$$P_2 = ft_{s_2} \left[1/6 \, (V_b I_c + V_c I_b) + 1/3 \, (V_b I_b + V_c I_c) \right] \tag{31}$$

$$P_3 = ft_{s_3} \left[1/6 \, (V_c I_a + V_a I_c) + 1/3 \, (V_c I_c + V_a I_a) \right] \tag{32}$$

where, as given above and defined in Figure 46:

$V_a = 3.0$ $\qquad$ $V_c = 1.5$

$I_b = 44.4 = I_c$

$V_b = 0.15$

$I_a = 0$

$t_{s_1} = 20 \, \mu$ sec.

$t_{s_2} = 10 \, \mu$ sec.

$t_{s_3} = 10 \, \mu$ sec.

P = Switching loss between points (a) and (b).

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* Formula from J. L. Jensen's Memo of April 8, 1957

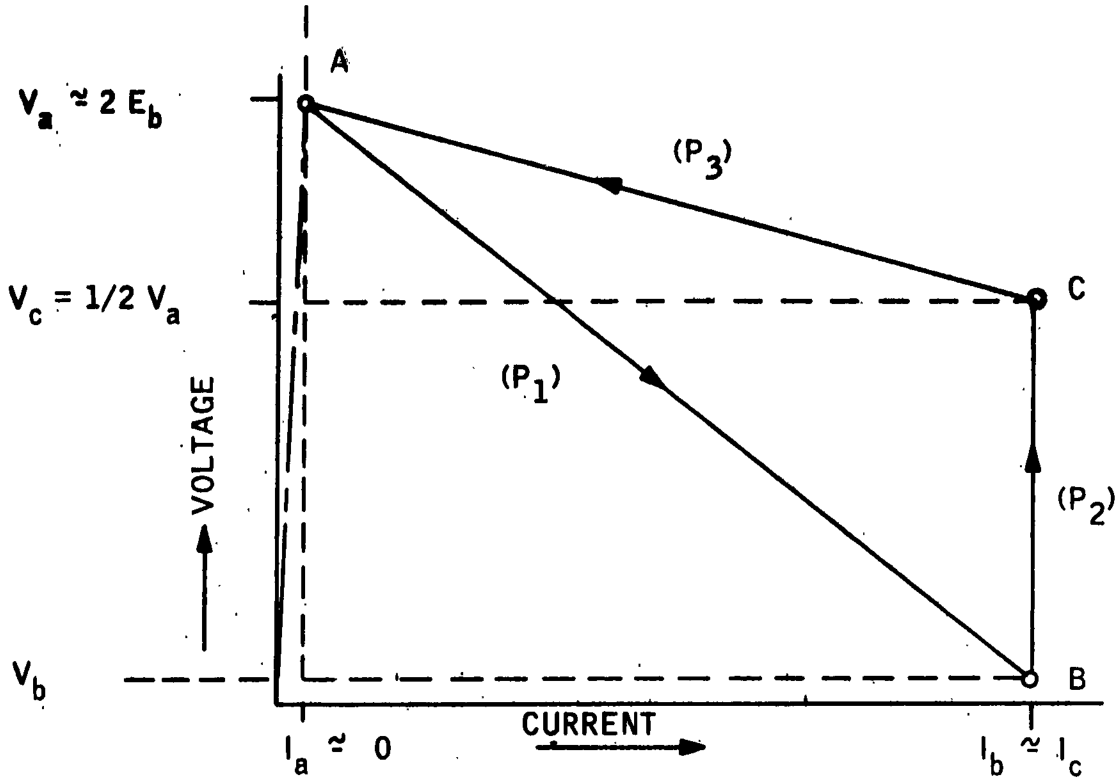

Figure 46 - ASSUMED SWITCHING CHARACTERISTICS

$$P_1 = f20\times10^{-6}\left[1/6(3.0\times44.4+0.15\times0)+1/3(3.0(0)+.15(44.4)\right]$$

$$= f20\times10^{-6}\left[22.2+2.22\right]$$

$$= f(20\times10^{-6})(24.4)$$

$$P_1 = 10^3(2\times10^{-5})(24.4) = .488 \text{ watts}$$

$$P_2 = f10\times10^{-6}\left[1/6(.15(44.4+1.5(44.4))+1/3(.15(44.4)+1.5(44.4)\right]$$

$$= f10\times10^{-6}\left[1.11+11.1+2.22+22.2\right]$$

$$= 10^3 \times 10\times10^{-6} \times 36.6 = .366 \text{ watts}$$

$$P_3 = f\times10\times10^{-6}\left[1/6(1.5(0)+3.0(44.4))+1/3(1.5(44.4)+3.0(0))\right]$$

$$= 10^3\times10\times10^{-6}\left[22.2+22.2\right]$$

$$= 10^3\times10\times10^{-6}\times(44.4) = .440 \text{ watts}$$

Thus total loss = $P_1 + P_2 + P_3 = P_T$

P_1 = .488 watts

P_2 = .366 watts

P_3 = .440 watts

P_T = 1.294 watts switching losses for each transistor

The switching losses for two transistors are 2x1.294 = 2.588 watts.

Rectifier Losses

If the output is 50 watts at 6.5 volts then the output current is $\dfrac{5.0 \text{ watts}}{6.5 \text{ volts}} = 7.7$ amps

If the rectifier voltage drop is 0.4 volt, the rectifier power loss $= 7.7 \times .4$
$= 3.08$ watts

$$\text{Rectifier efficiency} = \frac{\text{output}}{\text{output + losses}} = \frac{50}{50 + 3.08} = \frac{50}{53.08} = 94.2\% \qquad (33)$$

If the output power were taken at 28 volts the rectifier losses would be:

$I = \dfrac{50}{28} = 1.786$ amps. If the drop is .4 volts then $P = .4 \text{x} 1.786 = .715$ watts

$$\text{Rectifier efficiency} = \frac{50}{50 + .715} = 98.6\% \qquad (33)$$

If germanium were used for the 7.7 amp current then the forward drop (V_d) might be .25 volt and $V_d I = .25 \text{x} 7.7 = 1.93$ watts. Then efficiency $=$ $\dfrac{50}{51.93} = 96.4\%$. Thus the rectifier losses alone are appreciable and a 96% efficiency assumption is optimistic.

Drive Loss Calculations

With an assumed gain (h_{Fe}) of 40, estimate the required drive current.

$$I_b = \frac{I_c}{h_{Fe}} = \frac{44.4 \text{ amps}}{40} = 1.11 \text{ amps} \qquad (34)$$

Assuming 20% over drive this gives: 1.20x1.11 amps = 1.33 amps estimated drive. If the input voltage Veb = 1.2 volts, then the required drive is 1.2 volts x 1.33 amps or 1.6 watts.

If a resistance is incorporated in the feedback loop and is adjusted to dissipate half the drive power then the drive could go to 3.2 watts.

Summing the losses gives:

Switching losses	=	2.59 watts
Drive losses	=	3.20 watts
Drive & switching losses	=	5.79 watts

Other losses:

Transformer = (1-.94) 50	=	3.0 watts
Rectified filter = (1-.96) 50	=	2.0 watts
Total calculated losses	=	10.79 watts

Allowable loss = (input - output) = (66.7 - 50) = 16.7 watts

Allowable quiescent transistor loss = (16.7 - 10.79)

= 5.91 watts,

Assume leakage in "off" condition = 5 m. a.

Power = $5 \times 10^{-3} \times 3.0$ = .015 watts

Forward saturation loss = 5.90 watts

Maximum allowable forward saturation voltage $= \frac{\text{Power}}{I} = V_{ce}$ (Sat)

$$V_{ce}\ (\text{Sat}) = \frac{5.9 \text{ watts}}{44.44 \text{ amps}} \tag{35}$$

$= .133$ volt

Thus for a transistor having the above assumed characteristics the maximum V_{ce} (Saturation) = .133 volt at 44.44 amps. The calculated required characteristics of a transistor to achieve 75% efficiency in a transistor converter operating at 1.5 volts input are shown in Table I (see page 11).

B. THE TRANSISTOR APPROACH CONSIDERING A 1.0 VOLT SOURCE VOLTAGE AT FULL LOAD

The converter efficiency requirement is 75%.

The input collector current can be found from:

$$(28) \quad I_C = \frac{\text{Power Output}}{(\text{Source Voltage})\ \eta} = \frac{50 \text{ watts}}{1.0 \text{ volt} \times .75}$$

$$I_C = 66.7 \text{ amperes.}$$

Since the transistor current is considerably higher than in the 1.5 volt case it can be expected that a higher current gain would be desirable. Therefore assume a current gain of 60. The required base current is then:

$$(34) \quad I_b = \frac{I_C}{h_{Fe}} = \frac{66.7}{60} = 1.11 \text{ amps.}$$

For 20% over drive this becomes 1.2x1.11 = 1.33 amps.

If the emitter to base voltage is 1.2 volts the required drive power is 1.33 amps x 1.2 volts = 1.6 watts. If an additional 1.6 watts is lost in the drive circuitry, then the input drive power will be 3.2 watts which is the same as was assumed for the 1.5 volt case (with lower gain transistors).

The switching losses can again be calculated with equation (30), (31), and (32) assuming the switching characteristics shown in Figure 46 and the following assumptions:

$$V_a = 3.0$$

$$I_b = 66.7 = I_c$$

$$V_b = 0.15$$

$$V_c = 1.5$$

$I_a = 0$

$t_{s_1} = 20\ \mu$ sec.

$t_{s_2} = 10\ \mu$ sec.

$t_{s_3} = 10\ \mu$ sec.

(30) $P_1 = ft_{s_1}\left[1/6\,(V_aI_b + V_bI_a) + 1/3\,(V_aI_a + V_bI_b)\right]$

(31) $P_2 = ft_{s_2}\left[1/6\,(V_bI_c + V_cI_b) + 1/3\,(V_bI_b + V_cI_c)\right]$

(32) $P_3 = ft_{s_3}\left[1/6\,(V_cI_a + V_aI_c) + 1/3\,(V_cI_c + V_aI_a)\right]$

$P_1 = 10^3 \times 20 \times 10^{-6}\left[1/6\,(3.0\,(66.7) + .15\,(0)\,) + 1/3\,(3.0\,(0) + .15\,(66.7)\,)\right]$

$= 2 \times 10^{-2}\left[(33.3 + (3.33)\right]$

$= .732$ watt

$P_2 = 10^3 \times 10^{-5}\left[1/6\,(.15\,(66.7) + 1.5\,(66.7)\,) + 1/3\,(.15\,(66.7) + 1.5\,(66.7)\,)\right]$

$= 10^{-2}\,(\,(18.33) + (36.66)\,)$

$= .550$ watts

$P_3 = 10^3 \times 10^{-5}\left[1/6\,(1.5\,(0) + 3.0\,(66.7) + 1/3\,(1.5\,(66.7) + 3.0\,(0)\,)\right]$

$= 10^{-2}\,(33.3 + 3.33)$

$= .366$ watt

$$P_o = P_1 + P_2 + P_3$$

$$= .732 + .550 + .366$$

$$= 1.648 \text{ watts per transistor.}$$

For two transistors this becomes: $2P_o = 3.296$ watts or 3.30 watts.

Thus the switching losses for the 1.0 volt case are somewhat higher at 3.30 watts.

If the transformer efficiency is 94% the transformer losses are:

(36) $$P_{xfmr\ loss} = (1 - .94)\ 50$$
$$= 3 \text{ watts}$$

If the rectifier and filter efficiency is 96% these losses are:

(37) $$P_{Rect.\ and\ filter\ loss} = (1 - .96)\ 50$$
$$= 2 \text{ watts}$$

The total calculated losses are then:

Drive losses	=	3.20 watts
Switching losses =	=	3.30 watts
Transformer losses =	=	3.00 watts
Rectifier and filter losses	=	2.00 watts
Total		11.50 watts

The allowable losses are 66.7 - 50 = 16.7 watts.

The maximum quiescent losses can then be = 16.7-11.5
= 5.2 watts

If this is all consumed by V_{CE} (Sat.) . I_C then V_{CE} (Sat.) . I_C = 5.2 watts

or from equation (35) $V_{CE}\text{ (Sat.)} = \frac{5.2}{I_C} = \frac{5.2}{66.7}$

$= .078$ volt maximum

Thus the maximum V_{CE} Sat. voltage drop can be only .078 volt while the transistor is conducting 66.7 amperes. The required transistor parameters for 1.0 volt operation are shown in Table 2 (see Page 15).

Appendix A prepared by *John T. Lingle*
J. T. Lingle
Project Engineer

APPENDIX B

CALCULATION OF REQUIRED TUNNEL DIODE PARAMETERS

Sample calculations have been made to estimate the required tunnel diode characteristics to achieve 75% and 65% overall efficiencies with various input voltages. The results are shown in Tables IV, V, VI, VII, and VIII. The estimated curve for each case is shown in Figures 20, 21, and 22. By comparing the 75% efficiency requirements in Tables IV, V, and VI with the 65% efficiency requirements in Tables VII and VIII, it can be noted that the operating point ratios are more severe for the former. For about 1/2 volt input the I_1/I_2 operating point ratio is approximately three times greater for the 75% case than for the 65% case with other parameters nearly equal.

Larger ratios for the higher efficiency case can be noted by comparing the .25 volt input with the .368 volt input. Again, the current ratios are approximately three times as high for the 75% efficiency case. In addition, the voltage ratios are nearly twice as high. It must be pointed out that different assumptions were made in calculating these requirements. For the .25 volt input the I_1V_1 loss was assumed to be a major percentage of the total and this resulted in a low value of V_1 and V_p for the low voltage high current case. The value of V_p, is considerably lower than can be expected with the present state of the art. For the other cases V_p was taken at a higher value of .08 volt corresponding to a value obtainable in present commercial units. The higher values of V_p and V_1 necessitated higher supply voltages and valley voltages. Figure 20 shows the estimated curve for the lower voltage condition. Except for the rather low V_p and V_1 noted, this curve would represent a germanium tunnel diode. The curves of Figures 21 and 22 have higher source and valley voltages and these would require improved higher voltage materials such as gallium arsenide. Higher valley voltages shown in these curves might be theoretically possible although examination of typical specification sheets shows that present commercial devices have much lower valley voltages.

A. CALCULATIONS FOR A 0.25 VOLT SOURCE

The basic theoretical efficiency equation (see page 48) for the Tunnel Diode Converter of Figure 47 is:

$$\eta_o = \frac{1}{\left(1 + \frac{2 I_2}{I_1 - I_2}\right) \left(1 + \frac{2 V_1}{V_2 - V_1}\right)} \qquad (3)$$

where, as defined in Figure 19:

I_1 = near peak operating point

= $.9\, I_p$ (assumed)

I_2 = near valley operating point

I_2 = $1.1\, I_v$ (assumed)

V_1 = Voltage at I_1

V_2 = Voltage at I_2

Assume: Total circuit efficiency η = 75%
broken down as follows:

η_T = Transformer Efficiency = 94%

η_R = Rectifier and Filter Efficiency = 98%

η_S = Switching Efficiency = 99%

Thus total efficiency,

$$\eta = \eta_o \cdot \eta_T \cdot \eta_R \cdot \eta_S \qquad (38)$$

Since total efficiency = 75%,

$$.75\% = \eta_o \; (.94)\;(.98)\;(.99)$$

Required η_o is then:

$$\eta_o = \frac{.75}{.\;(.94)(.98)(.99)} = \frac{.75}{.912} = 82.3\%$$

Thus determine what tunnel diode characteristics are required to give a η_o of .823. For a 50 watt converter at 75% efficiency, the input power is $\frac{50}{.75}$ = 66.7 watts. Assuming that the supply voltage is .250 volts, the required input current is then $\frac{66.7}{.25}$ = 267 amperes.

This 267 amperes = $I_1 + I_2$ (as shown in Figure 47.)

Another equation for η_o includes the input voltage.

Thus: $\eta_o = \frac{\text{output}}{\text{input}}$ or

$$\eta_o = \frac{(I_1 - I_2)\,(E_b - V_1)}{(I_1 + I_2)\;E_b} = .823 \tag{39}$$

$$(I_1 - I_2)\,(E_b - V_1) = .823\;(I_1 + I_2)\,E_b = .823\;(267 \times .25) \tag{40}$$

$$= .823 \times 66.7$$

$$= 54.9$$

$$(I_1 - I_2)(.25 - V_1) = 54.9 \tag{41}$$

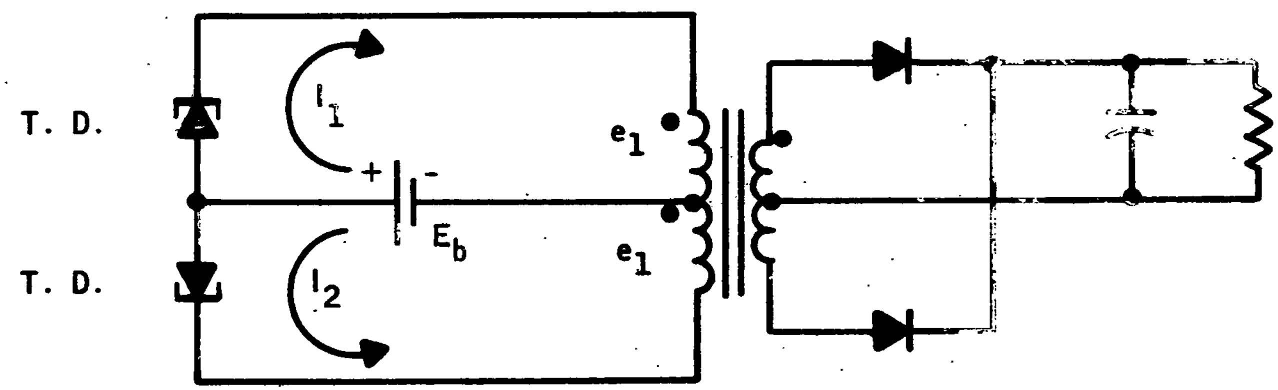

Figure 47 - TUNNEL DIODE CONVERTER

Also:

$$(I_1 - I_2) = \frac{54.9}{(.25 - V_1)} \tag{42}$$

and:

$$(I_1 + I_2) = 267 \tag{43}$$

Subtracting (43) from (42) gives:

$$-2I_2 = \left[\frac{54.9}{(.25 - V_1)} - 267\right] \quad \text{which reduces to:} \tag{44}$$

$$I_2 = \left[\frac{267 - \frac{54.9}{(.25 - V_1)}}{2}\right] \tag{45}$$

If a value is assumed for I_2, I_1, or V_1 the equation can be solved. Limiting cases can be established where either $I_2 = 0$ or $V_1 = 0$

if $V_1 = 0$

$$\text{Then:} \quad I_1 - I_2 = \frac{54.9}{.25} = 219.5 \text{ amperes} \tag{46}$$

$$\text{but:} \quad I_1 + I_2 = 267 \text{ amperes} \tag{47}$$

Subtracting equation (46) from (43) gives:

$2I_2 = 47.5$ amperes

$$I_2 \text{ maximum limit} = 23.75 \text{ amperes} \tag{48}$$

If $I_2 = 0$

then $I_1 + I_2 = 267$ amperes

or $I_1 = 267$ amperes

Therefore $(I_1 - I_2)(.25 - V_1) = 54.9$

$$(.25 - V_1) = \frac{54.9}{267}$$

$$V_1 = .25 - \frac{54.9}{267} = .25 - .2055$$

$$V_1 \text{ maximum limit } = .0445 \qquad (49)$$

In the practical case, most of the loss will probably be concentrated in the $I_1 V_1$ saturation voltage drop, therefore assume that the actual V_1 will be about .67 (V_1 maximum limit)

Therefore assume $V_1 = .67 \times .0445$

$= .0298$ volts

Going back to equation (41)

$$(I_1 - I_2)(.25 - .0298) = 54.9$$

$$I_1 - I_2 = \frac{54.9}{.2202}$$

$$I_1 - I_2 = 249 \text{ amperes} \qquad (50)$$

but:

$$I_1 + I_2 = 267 \text{ amperes} \qquad (43)$$

If $I_2 = 0$

then $I_1 + I_2 = 267$ amperes

or $I_1 = 267$ amperes

Therefore $(I_1 - I_2)(.25 - V_1) = 54.9$

$$(.25 - V_1) = \frac{54.9}{267}$$

$$V_1 = .25 - \frac{54.9}{267} = .25 - .2055$$

$$V_1 \text{ maximum limit} = .0445 \qquad (49)$$

In the practical case, most of the loss will probably be concentrated in the $I_1 V_1$ saturation voltage drop, therefore assume that the actual V_1 will be about .67 (V_1 maximum limit)

Therefore assume $V_1 = .67 \times .0445$

$= .0298$ volts

Going back to equation (41)

$$(I_1 - I_2)(.25 - .0298) = 54.9$$

$$I_1 - I_2 = \frac{54.9}{.2202}$$

$$I_1 - I_2 = 249 \text{ amperes} \qquad (50)$$

but:

$$I_1 + I_2 = 267 \text{ amperes} \qquad (43)$$

adding equations (50) and (43) gives: $2I_1 = 516$ amperes

$$I_1 = 258 \text{ amperes}$$

Therefore: $I_2 = 267 - I_1$

$$= 267 - 258$$

$$I_2 = 9 \text{ amperes}$$

Thus to obtain a device having 75% efficiency it must have the following operating points

$$I_1 = 258 \text{ amperes}$$

$$I_2 = 9 \text{ amperes}$$

$$V_1 = .0298 \text{ volts}$$

For practical operation it was assumed that $I_1 = .9\ I_p$.

Therefore $I_p = \dfrac{I_1}{.9} = \dfrac{258}{.9}$

$$I_p = 287 \text{ amperes}$$

also, I_2 was assumed to be equal to $1.1\ I_v$

Therefore $I_v = \dfrac{I_2}{1.1} = \dfrac{9}{1.1}$

$$I_v = 8.18 \text{ amperes.}$$

The peak current to valley current ratio for this device is:

$$I_p / I_v = \frac{287}{8.18} = 35.1$$

The operating point ratio $I_1 / I_2 = \dfrac{258}{9}$

$$= 28.7$$

The Tunnel Diode voltage at the "off" operating point is:

$$V_2 = 2\ Eb - V_1 \tag{51}$$

$$V_2 = 2\ (.250) - .0298$$

$$V_2 = .4702 \text{ volts}$$

Since operating point V_1 is not at the point of peak current and operating point V_2 is not at the point of valley current V_p and V_v will have to be estimated. The curves are not sine waves but a sine wave may be a reasonable approximation to the problem.

Thus if I_1 is assumed to be equal to:

$$I_1 = I_p \sin\ (\pi / 2 - \alpha)$$

where: α = angular displacement from peak point,

and $I_1 = .9\ I_p$

then: $\sin\ (\pi / 2 - \alpha) = .9$

and $(\pi / 2 - \alpha) = 64.2°$

also $\alpha = 25.8°$

For a sine wave function V_1 would then be:

$$V_1 = V_p \left(1 - \frac{\alpha}{\pi/2}\right) \tag{53}$$

$$= V_p \left(1 - \frac{25.8°)}{90°}\right.$$

$$= V_p \ (1 - .287)$$

$$V_1 = V_p \ (.713) \quad \text{or} \quad V_p = \frac{V_1}{.713}$$

If $V_1 = .0298$

$$\text{Then:} \quad V_p = \frac{V_1}{.713} = \frac{.0298}{.713}$$

$$V_p = .0418 \ \text{volt}$$

By a similar procedure, V_v can be estimated.

It might be assumed that a similar change would take place between V_2 and V_v. Inspection of typical tunnel diode curves shows that the curvature is considerably less in the valley point region than in the peak point region. Because of this it is estimated that the voltage difference between the valley operating point and the valley point will be about four times the difference between the peak operating point and the peak point for a comparable 10% change in current levels.

$$\text{Thus:} \quad V_v \simeq V_2 - 4 \ (V_p - V_1) \tag{54}$$

$$\simeq .4702 - 4 \ (.0418) - .0298)$$

$$\simeq .4702 - 4 \ (.012)$$

$$\simeq .4222 \ \text{volt}$$

Thus: V_v is estimated at .42 volt

$$\text{Ratio of } V_2/V_1 = \frac{.4702}{.0298}$$

$$V_2/V_1 = 15.77$$

$$\text{Ratio of } V_v/V_p = .42/.0418$$

$$V_v/V_p = 10$$

From this information the estimated required tunnel diode characteristics for satisfactory operation of a converter at 75% efficiency can be plotted. A plot of these estimated requirements is shown in Figure 20 (see page 54).

As a check on these estimates, we can insert our calculated values back into the η_o equation to determine if 82.3% can be obtained.

Thus substituting our operating points in equation (3) (see page 48).

$$\eta_o = \frac{1}{\left(1 + \frac{2\,I_2}{I_1 - I_2}\right)\left(1 + \frac{2\,V_1}{V_2 - V_1}\right)} \tag{3}$$

$$= \frac{1}{\left(1 + \frac{2\,(9)}{258-9}\right)\left(1 + \frac{2\,(.0298)}{.4702-.0298}\right)} = \frac{1}{\left(1 + \frac{18}{249}\right)\left(1 + \frac{.0596}{.4404}\right)}$$

$$= \frac{1}{(1.0723)\;(1.1353)} = \frac{1}{1.217}$$

$$\eta_o = 82.3\%$$

and thus the calculated results check with the original assumptions.

It may also be interesting to see what η_o would be if the device were operated at the calculated threshold using, I_p, V_p, I_v, and V_v.

Thus:

$$\eta_o = \frac{1}{\left(1 + \dfrac{2\,I_v}{I_p - I_v}\right)\left(1 + \dfrac{2\,V_p}{V_v - V_p}\right)} = \frac{1}{\left(1 + \dfrac{2\,(8.18)}{287 - 8.18}\right)\left(1 + \dfrac{2\,(.0418)}{.42\text{[illegible]} - .0418}\right)} \quad (3)$$

$$= \frac{1}{\left(1 + \dfrac{16.36}{278.82}\right)\left(1 + \dfrac{.0836}{.3794}\right)} = \frac{1}{(1.0587)\,(1.221)} \cong \frac{1}{1.292}$$

$$\eta_o = 77.4\%$$

and this is close to the other figure but appreciably lower. The primary reason for this is the higher voltage drop (V_p) during conduction. It can be noted that the ratio $\left[1 + \left(\dfrac{2\,V_p}{V_v - V_p}\right)\right]$ is greater

for this assumed threshold operation than for operation at operating points having current levels differing from the peak and valley points by 10%.

Some of the difference in results may be attributed to the assumption made that the tunnel diode curvatures approach that of a sine wave. If the formula for the tunnel diode curve were known it may be helpful to run maxima and minima calculations on operating point selections to determine where the optimum operating points are located. Certainly they are located reasonably close to the vicinity of the peak and valley points. Our estimates may be very close to the optimum.

Choosing current operating points 10% away from the peak and valley points appears to be necessary in building a practical inverter in order to obtain control over the operating frequency and prevent erratic operation at higher frequencies. The fact that the calculated efficiency came out higher at these operating points appears to be sufficiently optimistic, and further changes in operating points would probably not show any appreciable improvement.

TABLE IV

CALCULATED TUNNEL DIODE PARAMETERS REQUIRED TO CONSTRUCT A CONVERTER OPERATING AT 75% EFFICIENCY FROM A 0.25 VOLT SOURCE

Parameter	Value	Units
V_p	.0418	Volts
V_1	.0298	Volts
V_v	.42	Volts
V_2	.4702	Volts
I_p	287	Amps
I_1	258	Amps
I_v	8.18	Amps
I_2	9	Amps
$\frac{I_1}{I_2}$	28.7	
$\frac{I_p}{I_V}$	35.1	
$\frac{V_2}{V_1}$	15.77	
$\frac{V_v}{V_p}$	10.	

B. CALCULATIONS FOR HIGHER SOURCE VOLTAGES

Investigation shows that a peak voltage, V_p, of .080 volts is a reasonable value to expect for this parameter. Calculations have therefore been made on this value, and the valley voltage, V_v, and source voltage E_b, were increased to achieve the desired 75% efficiency.

Therefore assume:

$$V_p = .080 \text{ volt.}$$

Examination of current tunnel diode specifications (Table IX) shows that valley point voltages range from .300 to .335 for germanium and from .450 to .500 for silicon and gallium arsenide. It would be desirable to obtain higher valley voltages.

At this point several assumptions can be made. Two sets of calculations have been made, one set for each of the following two assumptions:

A. Assume that the voltage ratio $V_v/V_p = 11$.

B. Assume that the operating point ratio, (I_1/I_2) equals (V_2/V_1).

1. Calculations Based on Assumption A

Assume:

$$V_v/V_p = 11$$

$$V_p = .080 \text{ volt}$$

$$P_{(output)} = 50 \text{ Watts}$$

$$\eta_{overall} = .75$$

$$\eta_o = .825$$

$$P \text{ (input)} = 66.7 \text{ watts}$$

$$V_2 = V_v + 4\,(V_p - V_1)$$

$$V_1 = .713 \; V_p$$

$$I_1 = .9 \; I_p$$

$$I_2 = 1.1 \; I_V$$

Then:

$$V_v = 11 \; V_p$$

$$= 11\,(.08)$$

$$V_v = .88 \text{ volts}$$

Also:

$$V_1 = .713 \; V_p = .713\,(.08)$$

$$V_1 = .057 \text{ volt}$$

and,

$$V_2 = V_v + 4\,(V_p - V_1)$$

$$= .88 + 4\,(.080 - .057) = .88 + .092$$

$$V_2 = .972 \text{ volts.}$$

Writing the loop equations from Figure 47:

$$E_b - V_1 - e_1 = 0 \qquad (55)$$

$$E_b - V_2 + e_1 = 0 \qquad (56)$$

Adding (55) and (56) gives:

$$2E_b - V_1 = V_2 \quad \text{or} \tag{51}$$

$$2E_b = V_1 + V_2 \quad \text{and,}$$

$$E_b = \frac{V_1 + V_2}{2} \tag{57}$$

Substituting values in (57) gives:

$$E_b = \frac{V_1 + V_2}{2} = \frac{.057 + .972}{2}$$

$$E_b = .5145 \text{ volt}$$

The input current $(I_1 + I_2)$ can now be obtained from:

$$(I_1 + I_2) = \frac{P_{(input)}}{E_b}$$

$$= \frac{66.7 \text{ watts}}{.5145 \text{ volt}}$$

$$I_1 + I_2 = 129.6 \text{ amps.} \tag{59}$$

From the efficiency equation

$$\eta_o = \frac{(I_1 - I_2)}{(I_1 + I_2)} \frac{(E_b - V_1)}{E_b} = .823 \quad \text{or} \tag{39}$$

$$(I_1 - I_2) = \frac{.823\,(I_1 + I_2)\,E_b}{E_b - V_1}$$

Substituting values gives:

$$(I_1 - I_2) = \frac{.823\ (129.6)\ .5145}{.5145 - .057}$$

$$\therefore (I_1 - I_2) = 119.7 \text{ amps.} \tag{60}$$

Adding (59) and (60) gives:

$$2I_1 = 129.6 + 119.7$$

$$I_1 = \frac{129.6 + 119.7}{2} = \frac{249.3}{2}$$

$$I_1 = 124.65 \text{ amps}$$

Substituting back in (59)

$$124.65 + I_2 = 129.6 \qquad \text{or}$$

$$I_2 = 4.95 \text{ amps.}$$

From the original assumptions:

$$I_p = \frac{I_1}{.9} = \frac{124.65}{.9}$$

$$I_p = 138.5 \text{ amps}$$

$$I_v = \frac{I_2}{1.1} = \frac{4.95}{1.1}$$

$$I_v = 4.5 \text{ amps.}$$

These calculated parameters are tabulated in Table V. These results have also been plotted on Figure 21, page 54.

TABLE V

CALCULATED TUNNEL DIODE REQUIREMENTS FOR 75% EFFICIENCY BASED ON "ASSUMPTION A" AND A .515 VOLT SOURCE

Parameter	Value	Units
V_p	.080	Volts
V_1	.057	Volts
V_v	.880	Volts
V_2	.972	Volts
I_p	138.5	Amps
I_1	124.65	Amps
I_v	4.5	Amps
I_2	4.95	Amps
I_1/I_2	25.18	-----
I_p/I_v	30.75	-----
V_2/V_1	17.05	-----
V_v/V_p	11.0	-----

2. **Calculations Based on Assumption B**

Given:

$$V_2/V_1 = I_1/I_2$$

Then:

$$\frac{I_1 - I_2}{I_1 + I_2} = \frac{V_2 - V_1}{V_2 + V_1} = \sqrt{\eta_o} \tag{60}$$

for 75% efficiency $\eta_o = .823$

Therefore:

$$\frac{I_1 - I_2}{I_1 + I_2} = \sqrt{.823} = .907 \qquad \text{or}$$

$$I_1 - I_2 = .907\,(I_1 + I_2) \qquad \text{or}$$

$$I_1 = \frac{1.907}{.093} I_2$$

$$I_1 = 20.5 I_2 \tag{61}$$

Also, since the ratios were assumed equal

$$V_2 = 20.5\ V_1 \tag{62}$$

For assumed sine wave curvatures assume $V_p = .080$ volt, $I_1 = .9 I_p$, $V_2 = 1.1\ V_v$, and $V_1 = .713\ V_p$ as explained in the previous calculations.

Then:

$$V_1 = .713\ V_p$$

$$V_1 = .713 \times .08$$

$$= .057 \text{ volt}$$

$$V_2 = 20.5\, V_1 \tag{62}$$

$$= 20.5 \times .057$$

$$V_2 = 1.17 \text{ volts}$$

As in the previous calculations from (54) the valley voltage is assumed to have the following relationship.

$$V_v = V_2 - 4\,(V_p - V_1) \tag{63}$$

$$= 1.17 - 4\,(.080 - .057) = (1.17 - .092)$$

$$V_v = 1.078 \text{ volts.}$$

From the loop equation of Figure 47

$$V_2 = 2E_b - V_1 \tag{51}$$

$$1.17 = 2E_b - .057 \qquad \text{or}$$

$$2E_b = 1.227 \text{ volts} \qquad \text{and}$$

$$E_b = .6135 \text{ volt.}$$

Previous calculations (Item A) showed that the input power is 66.7 watts for 75% efficiency.

The input current $(I_1 + I_2)$ is then:

$$(I_1 + I_2) = \frac{P_{(input)}}{E_b} = \frac{66.7 \text{ watts}}{.6135 \text{ volt}} \tag{58}$$

$$I_1 + I_2 = 108.8 \text{ amps.} \tag{64}$$

As shown above:

$$\frac{I_1 - I_2}{I_1 + I_2} = .907 \quad \text{or}$$

$$I_1 - I_2 = .907\,(I_1 + I_2)$$

Substituting (64) gives:

$$I_1 - I_2 = .907\,(108.8) \tag{65}$$

Adding (64) and (65) gives:

$$2I_1 = 108.8 + .907\,(108.8)$$

$$I_1 = \frac{1.907\,(108.8)}{2}$$

$$= 103.74 \text{ amps.}$$

Substituting in (64) gives:

$$103.74 + I_2 = 108.8$$

$$I_2 = 5.06 \text{ amps.}$$

Since:

$$I_1 = .9I_p$$

$$I_p = \frac{I_1}{.9} = \frac{103.74}{.9}$$

$$I_p = 115.2 \text{ amps.}$$

Also:

$$I_2 = 1.1\, I_v$$

$$I_v = \frac{I_2}{1.1} = \frac{5.06}{1.1}$$

$$I_v = 4.6 \text{ amps.}$$

Thus, using the above assumptions required, tunnel diode parameters have been determined and are tabulated in Table VI below. From these values, a curve showing the required shape has been plotted on Figure 21, page 56.

TABLE VI

CALCULATED TUNNEL DIODE PARAMETERS REQUIRED TO CONSTRUCT A CONVERTER HAVING 75% EFFICIENCY OPERATING FROM A .614 VOLT SOURCE

Parameter	Value	Units
V_p	0.080	Volt
V_1	0.057	Volt
V_v	1.078	Volt
V_2	1.17	Volt
I_p	115.2	Amps
I_1	103.74	Amps
I_v	4.6	Amps
I_2	5.06	Amps
I_1/I_2	20.5	-----
I_p/I_v	25.03	-----
V_2/V_1	20.5	-----
V_v/V_p	13.47	-----

C. CALCULATION OF TUNNEL DIODE CHARACTERISTICS NECESSARY TO BUILD CONVERTERS HAVING 65% EFFICIENCY

To obtain 65% overall efficiency let:

$$\eta_o \cdot \eta_T \cdot \eta_R \cdot \eta_S = \eta = .65 \tag{66}$$

where:

η_T = Transformer efficiency = .94

η_R = Rectifier and filter efficiency = .98

η_S = Switching efficiency = .99

η_o = Basic operating point efficiency

η = Overall efficiency = .65

Thus rearranging (66):

$$\eta_o = \frac{\eta}{\eta_T \cdot \eta_R \cdot \eta_S} = \frac{.65}{(.94)(.98)(.99)} \tag{67}$$

$$\eta_o = .713$$

Thus with these assumptions the basic efficiency of the tunnel diode section η_o should be 71.3%.

The tunnel diode parameters can now be calculated using assumptions similar to "Assumption B" in the previous calculations. The following is assumed:

$$V_2/V_1 = I_1/I_2 \qquad \text{or}$$

$$\frac{I_1 - I_2}{I_1 + I_2} = \frac{V_2 - V_1}{V_2 + V_1} = \sqrt{\eta_o}$$

$$\eta = .65$$

$$\eta_o = .713$$

$$V_p = .080$$

$$I_p = I_1/.9$$

$$I_v = I_2/1.1$$

$$V_1 = .713\ V_p$$

$$V_v = V_2 - 4(V_p - V_1) \tag{63}$$

Substituting values:

$$V_1 = .713\ V_p = .713\ (.080)$$

$$V_1 = .057 \text{ volt}$$

Also:

$$\frac{V_2 - V_1}{V_2 + V_1} = \sqrt{\eta_o} = \sqrt{.713} \tag{68}$$

$$\frac{V_2 - V_1}{V_2 + V_1} = .845$$

$$V_2 - V_1 = .845\ V_2 + .845\ V_1$$

Which reduces to:

$$.155\ V_2 = 1.845\ V_1$$

$$V_2 = 11.9\ V_1$$

and by the same reasoning since

$$V_2/V_1 = I_1/I_2 \quad \text{then:}$$

$$I_1 = 11.9 I_2$$

Since $V_1 = .057$ volt, then

$$V_2 = 11.9 V_1 = 11.9\,(.057)$$

$$V_2 = .679 \text{ volt}$$

Equation (57) (see page B15) shows that the source voltage is:

$$E_b = \frac{V_2 + V_1}{2} \tag{57}$$

which gives:

$$E_b = \frac{V_2 + V_1}{2} = \frac{.679 + .057}{2}$$

$$E_b = .368 \text{ volt}$$

The input power is:

$$P_{(input)} = \frac{P_{(output)}}{\eta} = \frac{50 \text{ watts}}{.65} \tag{69}$$

$$= 76.9 \text{ watts}$$

The input current, $(I_1 + I_2)$ is given by:

$$(I_1 + I_2) = \frac{P_{(input)}}{E_b} \tag{58}$$

$$= \frac{76.9 \text{ watts}}{.368 \text{ volt}} \qquad \text{or}$$

$$(I_1 + I_2) = 209 \text{ amps.} \tag{70}$$

and since:

$$I_1 = 11.9 I_2$$

substituting in (70) gives:

$$11.9 I_2 + I_2 = 209 \text{ amps, or}$$

$$I_2 = \frac{209}{12.9}$$

$$I_2 = 16.2 \text{ amps.} \tag{71}$$

Then $I_1 = 209 - I_2$

$$= 209 - 16.2$$

$$I_1 = 192.8 \text{ amps.} \tag{72}$$

From the assumptions

$$I_p = I_1/.9 = \frac{192.8}{.9}$$

$$I_p = 214 \text{ amps.}$$

$$I_v = I_2/1.1 = \frac{16.2}{1.1}$$

$$I_v = 14.72 \text{ amps.}$$

$$V_v = V_2 - 4(V_p - V_1) = .679 - 4(.080 - .057) \tag{63}$$

$$V_v = .587 \text{ volt}$$

These calculated parameters are tabulated in Table VII. The required tunnel diode characteristic curve has been plotted on Figure 22 page 57.

TABLE VII

CALCULATED TUNNEL DIODE REQUIREMENTS FOR 65% EFFICIENCY AND A .368 VOLT SOURCE

Parameter	Value	Units
V_p	.080	Volt
V_1	.057	Volt
V_v	.587	Volt
V_2	.679	Volt
I_p	214	Amps
I_1	192.8	Amps
I_v	14.72	Amps
I_2	16.2	Amps
I_1/I_2	11.9	-----
I_p/I_v	14.53	-----
V_2/V_1	11.9	-----
V_v/V_p	7.34	-----

D. CALCULATION OF TUNNEL DIODE CHARACTERISTICS NECESSARY TO FABRICATE CONVERTERS OPERATING FROM A .500 VOLT SOURCE AT 65% EFFICIENCY

For another set of calculations, the following assumptions have been made:

Power output $= 50$ watts

Input voltage, $(E_b) = .500$ volt

$$V_p = .080 \text{ volt}$$
$$V_1 = .713\, V_p = .057 \text{ volt}$$
$$\eta = .65$$
$$\eta_o = .713$$
$$I_p = I_1/.9$$
$$V_v = V_2 - 4\,(V_p - V_1)$$
$$V_2 = 2E_b - V_1$$
$$P_{(input)} = 76.9 \text{ watts.}$$

Since:

$$V_2 = 2\,E_b - V_1 \tag{51}$$
$$= 2(.500) - .057$$
$$V_2 = .943 \text{ volt}$$

Then:

$$\eta_o = \left(\frac{I_1 - I_2}{I_1 + I_2}\right)\left(\frac{V_2 - V_1}{V_2 + V_1}\right) = .713 \tag{73}$$

$$\frac{I_1 - I_2}{I_1 + I_2} = \frac{.713}{\left(\frac{V_2 - V_1}{V_2 + V_1}\right)} \tag{74}$$

Substituting values gives:

$$\frac{I_1 - I_2}{I_1 + I_2} = \frac{.713}{\left(\frac{.943 - .057}{.943 + .057}\right)}$$

$$= \frac{.713}{\left(\frac{.886}{1.000}\right)}$$

$$\frac{I_1 - I_2}{I_1 + I_2} = .805 \qquad \text{or}$$

$$I_1 - I_2 = .805\, I_1 + .805\, I_2$$

$$.195\, I_1 = 1.805\, I_2$$

$$I_1 = 9.26\, I_2 \tag{75}$$

Also the input current $(I_1 + I_2)$ is:

$$I_1 + I_2 = \frac{P_{input}}{E_b} = \frac{76.9 \text{ watts}}{.50 \text{ volt}} \tag{58}$$

$$I_1 + I_2 = 153.8 \text{ amps} \tag{76}$$

Substituting (75) in (76) gives:

$$9.26\, I_2 + I_2 = 153.8 \text{ amps.}$$

$$I_2 = \frac{153.8}{10.26}$$

$$I_2 = 15.0 \text{ amps.} \quad (77)$$

Then:

$$I_1 = 153.8 - 15.0$$

$$I_1 = 138.8 \text{ amps} \quad (78)$$

From the initial assumptions:

$$I_p = I_1/.9 = \frac{138.8}{.9}$$

$$I_p = 154.1 \text{ amps.}$$

$$I_v = I_2/1.1 = \frac{15.0}{1.1}$$

$$I_v = 13.63 \text{ amps.}$$

$$V_v = V_2 - 4\,(V_p - V_1) = .943 - 4\,(.080 - .057) \quad (63)$$

$$V_v = .851 \text{ volt}$$

The calculated parameter requirements for the .500 volt input are tabulated in Table VIII. The required tunnel diode characteristic curve is plotted on Figure 22.

Appendix B prepared by John T. Lingle
J. T. Lingle
Project Engineer

TABLE VIII

CALCULATED TUNNEL DIODE REQUIREMENTS FOR CONVERTERS OPERATING FROM A .500 VOLT SOURCE AT 65% EFFICIENCY

Parameter	Value	Units
V_p	.080	Volt
V_1	.057	Volt
V_v	.851	Volt
V_2	.943	Volt
I_p	154.1	Amps
I_1	138.8	Amps
I_v	13.63	Amps
I_2	15.0	Amps
I_1/I_2	9.26	-----
I_p/I_v	11.3	-----
V_2/V_1	16.53	-----
V_v/V_p	10.63	-----

TABLE IX

REPRESENTATIVE PARAMETERS FOR TUNNEL DIODES PRESENTLY ON THE MARKET

Parameter	Minimum	Typical	Maximum	Material	Unit Rating
.01 Amp. Gallium Arsenide Unit*					
I_p (peak current)	9.0	10.0	11.0	Gallium	Milliamps
I_v (valley current)		0.5	.66	Arsenide	Milliamps
V_p (peak voltage)		0.10			Volt
V_v (valley voltage)		0.45			Volt
V_f (forward voltage at typical peak current)	.99	1.10	1.21	Gallium Arsenide	Volts
.1 Amp. Silicon Unit**					
I_p		100.00		Silicon	Milliamps
I_v			28.6		Milliamps
V_p		.075			Volt
V_v		0.47			Volt
V_f		0.78			Volt
I_p/I_v	3.5				
V_v/V_p		6.8		Silicon	

*From Texas Instrument's Bulletin No. D1-S 60322, March 1960.

**From Hoffman Electronics Corp., Preliminary Data Sheet Types HT 90 - HT 96, June 1962.

TABLE IX (Cont)

REPRESENTATIVE PARAMETERS FOR TUNNEL DIODES PRESENTLY ON THE MARKET

Parameter	Minimum	Typical	Maximum	Material	Unit Rating
10 Amp. Unit***					
I_p	9.0	10.0	11.0	Germanium	Amps.
I_v		2.0			Amps.
V_p	.090	.115	.145		Volt
V_v	.320	.345	.370		Volt
V_f	.515	.530	.565		Volt
I_p/I_v	8.0	10.0	12.0	Germanium	
200 Amp. Unit****					
I_p	180.00	200.00	220.00	Germanium	Amps.
I_v		25.0			Amps.
V_p			.110		Volt
V_v	.300		.370		Volt
V_f	.400		.480		Volt
I_p/I_v		8.0		Germanium	

***From RCA Tentative Data on High Current Germanium Tunnel Diode TD 191.

****From RCA Tentative Data on Extra High Current Germanium Tunnel Diode TD 226.

APPENDIX C

DERIVATION OF FORMULAS FOR BASIC EFFICIENCY CALCULATIONS ON CONVERTERS USING MAGNETORESISTIVE, SUPERCONDUCTIVE, PHOTORESISTIVE AND OTHER EFFECTS

The above approaches may be reduced to the push-pull equivalent circuit of Figure 48. In this circuit a DPDT switch alternately connects "high" and "low" resistances in series with each primary half of the push-pull output transformer. The switch and resistors represent the action of a switching transducer having "on" and "off" states for chopping d-c to a-c. The formulas for efficiency and optimum load resistance can be derived as a function of "off" and "on" transducer resistances. These derivations consider the transducer operating only at its quiescent "on" and "off" positions as shown below.

Assume that transformer losses and rectifier losses can be lumped into separate factors η_T and η_R respectively. The overall efficiency η will be given by:

$$\eta = (\eta_T) \ (\eta_R) \ (\eta_o) \tag{79}$$

where: η_o is the basic efficiency of the magnetoresistive converter neglecting the lumped losses as explained above.

The circuit is shown in Figure 48 and the operating points are shown in Figure 49.

The efficiency η_o will consider quiescent operation at the operating points (P_1), (P_2) and will not consider the switching losses incurred when switching from one point to the other.

For simplification assume $N_{1_A} = N_{1_B} = N_2$,

then the winding induced voltage e_1 is the same on each N_1 and N_2. Since I_2 opposes I_1, the power transformed to the secondary is:

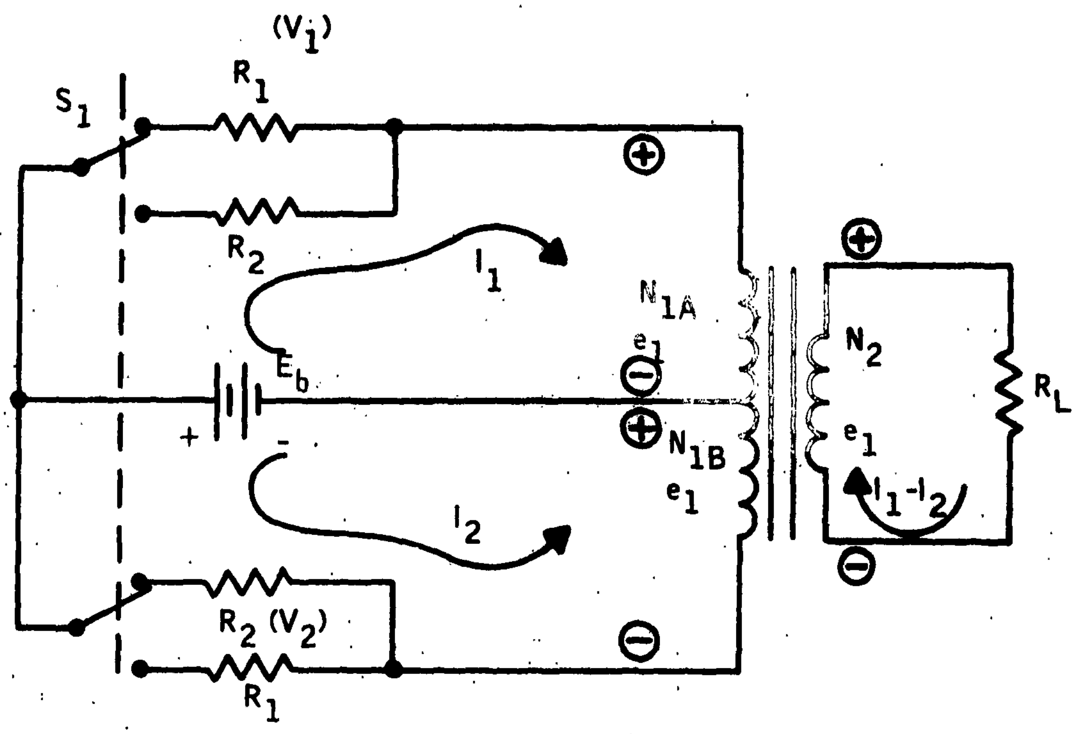

Figure 48 - BASIC CIRCUIT

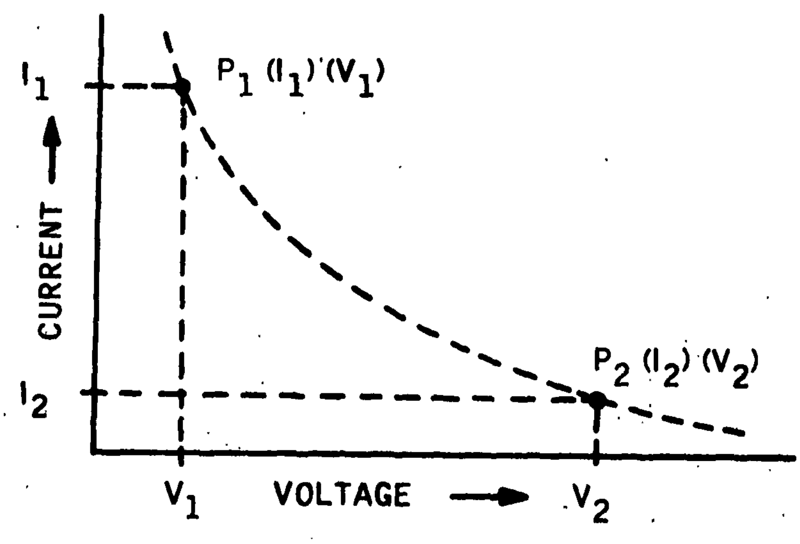

Figure 49 - OPERATING POINTS

$$P_{out} = (I_1 - I_2)\, e_1 \tag{80}$$

The voltage drop V_1 across magnetoresistor R_1 is:

$$V_1 = I_1 R_1 \tag{81}$$

Also the voltage drop V_2 across magnetoresistor R_2 is:

$$V_2 = I_2 \; R_2 \tag{82}$$

The loop equations can be written as follows:

$$E_b - I_1 R_1 - e_1 = 0 \tag{83}$$

$$E_b - I_2 R_2 + e_1 = 0 \tag{84}$$

subtracting (84) from (83) gives:

$$-2e_1 - I_1 R_1 + I_2 R_2 = 0 \qquad \text{or}$$

$$-2e_1 = I_1 R_1 - I_2 R_2 \qquad \text{or}$$

$$e_1 = \frac{I_2 R_2 - I_1 R_1}{2} \tag{85}$$

Also since $V_2 = I_2 R_2$ and $V_1 = I_1 R_1$, (81) and (82) then:

$$e_1 = \frac{V_2 - V_1}{2} \tag{86}$$

Since (80) gives the output as $(I_1 - I_2)\, e_1$, substituting (86) produces:

$$(I_1 - I_2) \frac{(V_2 - V_1)}{2} = P_{out} \tag{87}$$

The power input to this converter is

$$P_{(input)} = E_b\ (I_1 + I_2). \tag{88}$$

Adding equations (83) and (84) gives:

$$2\,E_b - I_1R_1 - I_2R_2 - e_1 + e_1 = 0 \quad \text{or}$$

$$2\,E_b = I_1R_1 + I_2R_2 \tag{89}$$

Since $V_1 = I_1R_1$, $V_2 = I_2R_2$; substituting (81) and (82) in (89) gives:

$$2\,E_b = V_1 + V_2 \quad \text{or}$$

$$E_b = \frac{V_1 + V_2}{2} \qquad \text{(This is identical to (57))} \tag{90}$$

Thus the power input is found by substituting (90) in (88) producing:

$$P_{(in)} = E_b\ (I_1 + I_2) = \frac{(V_1 + V_2)}{2}\ (I_1 + I_2) \tag{91}$$

Efficiency is defined as:

$$\eta_o = \frac{P_{(out)}}{P_{(in)}} \tag{92}$$

Substituting (87) and (91) in (92) gives:

$$\eta_o = \frac{P_{(out)}}{P_{(in)}} = \frac{(I_1 - I_2)}{(I_1 + I_2)} \frac{\left(\frac{V_2 - V_1}{2}\right)}{\left(\frac{V_2 + V_1}{2}\right)} \quad \text{or}$$

$$\eta_o = \frac{(I_1 - I_2)\ \ (V_2 - V_1)}{(I_1 + I_2)\ \ (V_2 + V_1)} \tag{93}$$

It can be seen that this is the same equation as (3) and (73) used by Hanrahan*. This can be divided into two parts

$$\eta_o = \left[\frac{I_1 - I_2}{I_1 + I_2}\right] \cdot \left[\frac{V_2 - V_1}{V_2 + V_1}\right] \tag{93}$$

consider $\left[\frac{I_1 - I_2}{I_1 + I_2}\right]$ and express in terms of R_1, R_2 and R_L

From (81),

$$I_1 = \frac{V_1}{R_1} = \frac{E_b - e_1}{R_1} \tag{94}$$

Now assume that there is a synthetic resistance, R_f, reflected into N_{1_A} defined such that:

$$I_1 R_f = e_1. \tag{95}$$

Then writing the loop equation for I_1

$$E_b - I_1R_1 - I_1R_f = 0 \qquad \text{or}$$

$$E_b = I_1 \ (R_1 + R_f)$$

$$\text{and } I_1 = \frac{E_b}{R_1 + R_f} \tag{96}$$

Writing the loop equation for I_2 gives:

$$E_b - I_2R_2 + e_1 = 0 \tag{97}$$

Substituting (95) in (97) gives:

$$E_b - I_2R_2 + I_1R_f = 0$$

$$E_b + I_1R_f = I_2R_2 \qquad \text{or}$$

* D. J. Hanrahan, "Analysis of Tunnel Diode Converter Performance", IRE TRANSACTIONS ON ELECTRON DEVICES, July 1962.

$$I_2 = \frac{E_b + I_1 R_f}{R_2} \tag{98}$$

but since $I_1 = \dfrac{E_b}{R_1 + R_f}$, substituting (96) in (98) gives:

$$I_2 = \frac{E_b + \dfrac{E_b R_f}{(R_1 + R_f)}}{R_2} \tag{99}$$

The resistance values can now be substituted in the quantity $\left[\dfrac{I_1 - I_2}{I_1 + I_2}\right]$ by utilizing equations (96) and (99) producing:

$$\left[\frac{I_1 - I_2}{I_1 + I_2}\right] = \frac{\left[\dfrac{E_b}{R_1 + R_f}\right] - \left(\dfrac{E_b}{R_2}\right) - \left[\dfrac{E_b R_f}{R_2 (R_1 + R_f)}\right]}{\left[\left(\dfrac{E_b}{R_1 + R_f}\right) + \left(\dfrac{E_b}{R_2}\right) + \left(\dfrac{E_b R_f}{R_2 (R_1 + R_f)}\right)\right]}$$

it can be noted that E_b cancels out leaving

$$\left[\frac{I_1 - I_2}{I_1 + I_2}\right] = \frac{\left(\dfrac{1}{R_1 + R_f}\right) - \left(\dfrac{1}{R_2}\right) - \left(\dfrac{R_f}{R_2 (R_1 + R_f)}\right)}{\left(\dfrac{1}{R_1 + R_f}\right) + \left(\dfrac{1}{R_2}\right) + \left(\dfrac{R_f}{R_2 (R_1 + R_f)}\right)}$$

placing under common denominators gives:

$$\left[\frac{I_1 - I_2}{I_1 + I_2}\right] = \frac{\left[\left(\dfrac{R_2}{R_2 (R_1 + R_f)}\right) - \left(\dfrac{(R_1 + R_f)}{R_2 (R_1 + R_f)}\right) - \left(\dfrac{R_f}{R_2 (R_1 + R_f)}\right)\right]}{\left[\left(\dfrac{R_2}{R_2 (R_1 + R_f)}\right) + \left(\dfrac{R_1 + R_f}{R_2 (R_1 + R_f)}\right) + \left(\dfrac{R_f}{R_2 (R_1 + R_f)}\right)\right]}$$

Note that the denominator $\left[R_2 (R_1 + R_f)\right]$ cancels giving:

$$\left[\frac{I_1 - I_2}{I_1 + I_2}\right] = \left[\frac{(R_2 - R_1 - R_f - R_f)}{(R_2 + R_1 + R_f + R_f)}\right]$$

$$\left[\frac{I_1 - I_2}{I_1 + I_2}\right] = \left[\frac{(R_2 - R_1 - 2R_f)}{(R_2 + R_1 + 2R_f)}\right] \qquad (100)$$

Now substitute values to express $\left[\frac{V_2 - V_1}{V_2 + V_1}\right]$ in terms of R_1, R_2, and R_f.

Substituting (81) and (82) in this quantity gives:

$$\frac{(V_2 - V_1)}{(V_2 + V_1)} = \left(\frac{I_2R_2 - I_1R_1}{I_2R_2 + I_1R_1}\right) \qquad (101)$$

but from (98) $I_2 = \left(\frac{E_b + I_1R_f}{R_2}\right)$ and

and from (96) $I_1 = \frac{E_b}{R_1 + R_f}$

Substituting (98) and (96) in (101) gives:

$$\left[\frac{V_2 - V_1}{V_2 + V_1}\right] = \frac{\left[\frac{E_b + I_1R_f}{R_2}\right]R_2 - \left[\frac{E_b R_1}{R_1 + R_f}\right]}{\left[\frac{E_b + I_1R_f}{R_2}\right]R_2 + \left[\frac{E_b R_1}{R_1 + R_f}\right]}$$

and this reduces to:

$$\left[\frac{V_2 - V_1}{V_2 + V_1}\right] = \frac{E_b + I_1 R_f - \dfrac{E_b R_1}{(R_1 + R_f)}}{E_b + I_1 R_f + \dfrac{E_b R_1}{(R_1 + R_f)}} \tag{102}$$

Since $I_1 = \dfrac{E_b}{R_1 + R_f}$, then:

$$\left[\frac{V_2 - V_1}{V_2 + V_1}\right] = \frac{\dfrac{E_b + E_b R_f}{(R_1 + R_f)} - \dfrac{E_b R_1}{(R_1 + R_f)}}{\dfrac{E_b + E_b R_f}{(R_1 + R_f)} + \dfrac{E_b R_1}{(R_1 + R_f)}}$$

it can be noted that E_b cancels out.

$$\left[\frac{V_2 - V_1}{V_2 + V_1}\right] = \frac{\dfrac{1 + Rf}{(R_1 + R_f)} - \dfrac{R_1}{(R_1 + R_f)}}{\dfrac{1 + R_f}{(R_1 + R_f)} + \dfrac{R_1}{(R_1 + R_f)}}$$

Placing this under a common denominator gives:

$$\left[\frac{V_2 - V_1}{V_2 + V_1}\right] = \frac{\dfrac{R_1 + R_f + R_f - R_1}{(R_1 + R_f)}}{\dfrac{R_1 + R_f + R_f + R_1}{(R_1 + R_f)}}$$ and this reduces to:

$$\left[\frac{V_2 - V_1}{V_2 + V_1}\right] = \left[\frac{2R_f}{2R_1 + 2R_f}\right] = \left[\frac{R_f}{R_1 + R_f}\right]$$

$$\left[\frac{V_2 - V_1}{V_2 + V_1}\right] = \left[\frac{R_f}{R_1 + R_f}\right] \tag{103}$$

Substituting (100) and (103) into (93), these values now give:

$$\eta_o = \left[\frac{I_1 - I_2}{I_1 + I_2}\right] \quad \left[\frac{V_2 - V_1}{V_2 + V_1}\right] \tag{93}$$

$$\eta_o = \left[\frac{R_2 - R_1 - 2R_f}{R_2 + R_1 + 2R_f}\right]\left[\frac{R_f}{R_1 + R_f}\right] \tag{104}$$

Equation (104) appears to be in reasonably simple form.

The power output of this device is

$$P_{(out)} = (I_1 - I_2)^2 R_L \text{ where } R_L = \text{actual} \tag{105}$$

load resistance if the turns ratio is 1:1.

Because of our assumptions the power output also equals:

$$P_{(out)} = e_1^2 / R_L \tag{106}$$

Therefore equating (105) and (106) gives:

$$(I_1 - I_2)^2 R_L = \frac{e_1^2}{R_L}, \text{ But by Definition} \tag{107}$$

$$e_1 = I_1 R_f \tag{95}$$

Substituting (95) in (107) gives:

$$(I_1 - I_2)^2 R_L = \frac{I_1^2 R_f^2}{R_L} \quad \text{or} \tag{108}$$

$$(I_1 - I_2)^2 R_L^2 = I_1^2 R_f^2 \quad \text{and}$$

$$R_f^2 = \frac{(I_1 - I_2)^2 R_L^2}{I_1^2}$$

Taking square roots and assuming the positive root gives:

$$R_f = \frac{(I_1 - I_2) R_L}{I_1} \tag{109}$$

but substituting (96), $\left[I_1 = \frac{E_b}{R_1 + R_f} \right]$

and (98) $\left[I_2 = \frac{E_b + \frac{E_b R_f}{R_1 + R_f}}{R_2} \right]$ into (109) gives:

$$R_f = \left[\frac{\left(\frac{E_b}{R_1 + R_f}\right) - \left(\frac{E_b}{R_2}\right) - \left(\frac{E_b R_f}{R_2 (R_1 + R_f)}\right) \quad R_L}{\left(\frac{E_b}{R_1 + R_f}\right)} \right] \tag{110}$$

E_b cancels and placing under common denominators gives:

$$R_f = \frac{\left[\frac{(R_2 - (R_1 + R_f) - R_f}{R_2 (R_1 + R_f)} \right] R_L}{\frac{1}{R_1 + R_f}} \quad \text{or}$$

$$R_f = \frac{(R_2 - R_1 - R_f - R_f) \quad R_L}{R_2} = \frac{(R_2 - R_1 - 2R_f) \quad R_L}{R_2} \quad \text{or}$$

$$R_2 R_f = (R_2 - R_1 - 2R_f) \quad R_L$$

Solving for R_L gives:

$$R_L = \frac{R_2 R_f}{R_2 - R_1 - 2R_f} \tag{111}$$

Equation (111) may also be solved for R_f giving:

$$R_2 R_f = (R_2 - R_1) \ R_L - 2R_f R_L$$

$$R_f (R_2 + 2R_L) = R_L (R_2 - R_1)$$

$$R_f = \left[\frac{R_L \quad (R_2 - R_1)}{R_2 + 2R_L}\right] \quad \text{(assuming 1:1 turn ratios)} \quad \text{for other ratios multiply by} \left[\frac{N_1}{N_2}\right]^2 \tag{112}$$

This expression could then be substituted for R_f if desired, giving:

$$\eta_o = \left[\frac{R_2 - R_1 - 2\left(\frac{R_L(R_2 - R_1)}{R_2 + 2R_L}\right)}{R_2 + R_1 + 2\left(\frac{R_L(R_2 - R_1)}{R_2 + 2R_L}\right)}\right]\left[\frac{\frac{R_L(R_2 - R_1)}{R_2 + 2R_L}}{R_1 + \frac{R_L(R_2 - R_1)}{R_2 + 2R_L}}\right] \tag{113}$$

using the actual value of R_L makes the equation (113) more complex than (104). Use equation (104) for calculations.

Equation (79) shows that the overall efficiency $\eta = (\eta_o)(\eta_T)(\eta_R)$.

This does not include switching losses. If the switching losses are considered, another lumped factor can be used for these. Let η_s = switching efficiency defined as:

$$\eta_s = \frac{\text{output}}{\text{output + switching losses}} \tag{114}$$

Switching losses can probably be obtained using equation (30)*.

---

* Jensen's formula for switching loss (transient)

$$P = ft_s \left[1/6\,(V_a I_b + V_b I_a) + 1/3\,(V_a I_a + V_b I_b)\right] \tag{30}$$

where: (Symbols are defined in Appendix A, Page A2 .)

The overall efficiency will be

$$\eta = \eta_o \, \eta_T \, \eta_R \, \eta_S \simeq 75\% \text{ (desired)} \tag{115}$$

assume $\eta_T = 94\%$

$\eta_R = 96\%$ or 98%

then required.

$$\eta_o = \frac{.75}{.94 \times .96 \times \eta_S} \tag{116}$$

The above equation (113) for η_o is more complex because the load resistance R_L or R_f is included. Since the required load is known, the quantity $(I_1 - I_2)$ is a known factor. For any given input, voltage E_b is known which should enable one to estimate V_1 and V_2. Also the quantity $(V_2 - V_1)$ should be a known factor since $e_1 = \frac{V_2 - V_1}{2}$. The first approximations should be made using (93)

$$\eta_o = \left[\frac{I_1 - I_2}{I_1 + I_2}\right] \left[\frac{V_2 - V_1}{V_2 + V_1}\right] \quad \text{since these values can be estimated.} \tag{93}$$

Using this information, preliminary estimates can be made for the values of R_1, R_2, R_L, and R_f.

A. DETERMINATION OF OPTIMUM LOAD FOR MAXIMUM EFFICIENCY

The efficiency equation is:

$$\eta_o = \left[\frac{(R_2 - R_1 - 2R_f)}{(R_2 + R_1 + 2R_f)}\right]\left[\frac{R_f}{R_1 + R_f}\right] \quad (117)$$

Let η_o = y, and R_f = x , so that:

$$y = \left[\frac{R_2 - R_1 - 2x}{R_2 + R_1 + 2x}\right]\left[\frac{x}{R_1 + x}\right] \quad (118)$$

To find the maximum and minimum points where the slope is zero, differentiate and set dy/dx = 0.

Take $\left[\frac{R_2 - R_1 - 2x}{R_2 + R_1 + 2x}\right]$ and differentiate first

using $d\left(\frac{u}{v}\right) = \frac{vdu - udv}{v^2}$. Thus for this quantity $[\alpha]$ = (119)

$\left[\frac{R_2 - R_1 - 2x}{R_2 + R_1 + 2x}\right]$:

$$\frac{d[\alpha]}{dx} = \frac{[R_2 + R_1 + 2x](-2) - [R_2 - R_1 - 2x](2)}{[R_2 + R_1 + 2x]^2} \quad (120)$$

Next, differentiate $\left[\frac{x}{R_1 + x}\right]$ and for this,

$$\frac{d\left[\frac{x}{R_1 + x}\right]}{dy} = \left[\frac{[R_1 + x](1) - x\ (1)}{[R_1 + x]^2}\right] \quad (121)$$

Now use: (122)

$$d\,(uv) = u\,dv + v\,du$$

So: (123)

$$\frac{dy}{dx} = \left[\frac{(R_2-R_1-2x)}{(R_2+R_1+2x)}\right]\left[\frac{(R_1+x-x)}{(R_1+x)^2}\right] + \left[\frac{x}{R_1+x}\right]\left[\frac{-2(R_2+R_1+2x)-2\,(R_2-R_1-2\,x)}{(R_2+R_1+2x)^2}\right]$$

$$\frac{dy}{dx} = \left[\frac{(R_2-R_1-2x)\,R_1}{(R_2+R_1+2x)(R_1+x)^2}\right] + x\left[\frac{-2\,R_2-2R_1-4x-2R_2+2R_1+4x}{(R_1+x)(R_2+R_1+2x)^2}\right] \quad \text{or}$$

$$\frac{dy}{dx} = \left[\frac{(R_2 - R_1 - 2x)\,R_1}{(R_2 + R_1 + 2x)\,(R_1 + x)^2} - \frac{4R_2x}{(R_1 + x)\,(R_2 + R_1 + 2x)^2}\right]$$

Setting $\frac{dy}{dx} = 0$ gives:

$$\frac{(R_2 - R_1 - 2x)R_1}{(R_2 + R_1 + 2x)\,(R_1 + x)^2} - \frac{4R_2x}{(R_1 + x)\,(R_2 + R_1 + 2x)^2} = 0 \qquad (124)$$

or

$$\frac{(R_2 - R_1 - 2x)R_1}{(R_2 + R_1 + 2x)\,(R_1 + x)^2} = \frac{4R_2x}{(R_1 + x)\,(R_2 + R_1 + 2x)^2}$$

Multiply both sides by $(R_2 + R_1 + 2x)\,(R_1 + x)$ to give:

$$\frac{(R_2 - R_1 - 2x)R_1}{R_1 + x} = \frac{4R_2x}{(R_2 + R_1 + 2x)}$$

Cross multiply to give:

$$\left[R_2 - (R_1 + 2x)\right](R_2 + R_1 + 2x)R_1 = 4R_2x(R_1 + x) \qquad \text{or}$$

$$\left[R_2^2 - (R_1 + 2x)^2\right]R_1 = 4R_2R_1x + 4R_2x^2$$

Expanding,

$$R_2^2R_1 - R_1^3 - 4R_1^2x - 4R_1x^2 = 4R_2R_1x + 4R_2x^2$$

Collecting terms and transposing gives:

$$-4x^2(R_1 + R_2) - 4x\,(R_1^2 + R_2R_1) - R_1^3 + R_2^2R_1 = 0$$

Multiply by (-1):

$$4x^2(R_1 + R_2) + 4x(R_1^2 + R_2R_1) + R_1^3 - R_2^2R_1 = 0 \qquad (125)$$

Dividing by $(R_1 + R_2)$ gives:

$$4x^2 + 4xR_1 + R_1(R_1 - R_2) = 0 \qquad (126)$$

Solving for x with the quadratic formula gives:

$$x = \frac{-R_1 \pm\sqrt{R_1 R_2}}{2} \qquad (127)$$

Inspection indicates that the positive square root yields the desired answer. The fact that this is a maximum point has not been rigorously proven but this is assumed to be the case. Substituting R_f for x now gives the desired formula for the optimum R_f.

$$R_{f_{(optimum)}} = \frac{-R_1 \pm\sqrt{R_1 R_2}}{2} \qquad (128)$$

B. CONCLUSIONS

Calculations have been made using the above formulas to obtain optimum R_f and optimum η_o for a transducer impedance range from 1:1 to 2000:1. These calculated results have been plotted on Figure 28. The optimum value of R_L referenced to the primary has also been plotted for this impedance range. The resistance curves indicate what value R_1 should be multiplied by in order to arrive at the optimum load resistance.

The value of resistors R_1, R_2 will also be determined by the input voltage and required power output. It can be noted that the efficiency is zero for an R_2/R_1 ratio of 1.0. Ratios of 10, 100, and 400 give theoretical basic efficiencies of 26.95%, 67.0%, and 82.7% respectively. With higher impedance ratios, the curve asymptotically approaches 100%. It can be noted that high ratios above 400 will be required to build the required converter. The other lumped parameter efficiencies η_T, η_r, and η_S must be considered in making these estimates. The information on this curve should be useful in estimating transducer requirements. This curve should also be useful in estimating what efficiencies can be achieved using present state-of-the-art materials.

Appendix C
Prepared by:

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APPENDIX D

CALCULATIONS ON THE SUPERCONDUCTIVE APPROACH

A. CALCULATION OF REFRIGERATION POWER REQUIRED

The heat conducted into the cryostat through the input leads and walls and the heat generated inside the cryostat by the transducer must be removed by some form of refrigeration. This energy must be removed in order to maintain the cryotron at operating temperature. The minimum amount of power required to extract energy from a low temperature region and reject it at a higher temperature region can be found from Carnot's principle thus:

$$P = W\left[\frac{T_2}{T_1} - 1\right] \tag{130}$$

where:

P = Minimum refrigeration power required
W = Dissipation inside the cryostat
T_1 = Low temperature = $4.2^{\circ}K$
T_2 = High temperature assumed to be $290^{\circ}K$

For each watt of internal dissipation the following refrigeration power is required:

$$P = (1)\left[\frac{290}{4.2} - 1\right]$$

$$= \quad 69.1 - 1$$

$$P = \quad 68.1 \text{ watts minimum.}$$

In a practical case it would probably be much greater than this because the efficiency of the refrigeration system is not 100%. This refrigeration power requirement might be obtained from the device output by the utilization of some form of electromechanical refrigerator and combinations of Peltier cooling effects might be feasible. The evaporation of a quantity of liquid helium is involved during some phase of the process. If a refrigerator is used, the helium supply is reliquified and reused. The required refrigerator would add considerable weight and complexity. Another approach consists of carrying a large supply of liquid helium in a super-insulated vessel.

B. CALCULATION OF POWER LOSS DUE TO HEAT CONDUCTION THROUGH THE INPUT LEADS

Refrigeration power must be supplied to remove each watt of lead heat flow to maintain the device at operating temperature. It has been shown that for an optimum copper lead pair * 0.084 watt of heat energy will be conducted into the device for each ampere carried. The optimum refrigeration power required is then:

$$P_{RF} = .084 \frac{\text{watt}}{\text{ampere}} \times 68.1 \frac{\text{watts (refrigeration)}}{\text{watt (conducted heat)}} \tag{131}$$

P_{RF} (optimum) = 5.72 watts/ampere

If half of the .084 watt heat flow is assumed to be caused by I^2R loss and the other half by the temperature difference, the assumed power dissipation in the lead would be approximately .04 watt. The total loss per ampere (P_L) would then be 5.72 plus .04 watt or (P_L) equals 5.76 watts per ampere.

A theoretical efficiency factor η_{HRf} can be defined assuming that refrigeration power will be obtained from the converted device output at 100% efficiency.

Considering the header lead heat flow factor and refrigeration power required to compensate for it as a lumped parameter, η_{HRf}, the theoretical maximum efficiency due to this factor would be:

$$\eta_{HRf} = \frac{\text{input - (header heat flow + compensating refrigeration)}}{\text{input}} \quad \text{or}$$

$$\eta_{HRf} = \frac{E_b I_o - I_o P_L}{E_b I_o} \tag{132}$$

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* R. McFee "Optimum Input Leads for Cryogenic Apparatus" - Review of Scientific Instruments - Vol. 30 Feb. 1959, p. 98.

where: E_b = Input voltage
I_o = Input current
P_L = Heat flow through header leads plus compensating refrigeration required per ampere or 5.76 $\frac{\text{watts}}{\text{ampere}}$.

Equation (132) can be written as:

$$\eta_{HRf} (E_b I_o) = E_b I_o - I_o P_L$$

It can be noted that I_o cancels out leaving:

$$\eta_{HRf} E_b = E_b - P_L \tag{133}$$

This can also be written as:

$$E_b = \frac{P_L}{1 - \eta_{HRf}} \tag{134}$$

To construct a practical device it would be desirable to make $\eta_{HRf} \geq 90\%$. The minimum source voltage can be determined by substituting these values in equation (134) as follows:

$$E_{b\,(Min)} \geq \frac{P_L}{1 - \eta_{HRf}} \geq \frac{5.76}{1 - .90} \quad \text{or}$$

$$E_{b\,(Min)} \geq 57.6 \text{ volts.}$$

This voltage is very high when compared with our anticipated source which ranges between 0.1 and 1.5 volts. The minimum source voltage for $\eta_{HRf} =$ 75% can also be obtained by substituting values in (134):

$$E_{b\,(Min.)} \geq \frac{5.76}{1 - .75} \quad \text{or}$$

$$E_{b\,(Min.)} \geq 23.1 \text{ volts}$$

This voltage is also too high. Another interesting E_b parameter is the point of zero efficiency. Equation (134) shows that if $\eta_{HRf} = O$, then $E_b = P_L$ or $E_b = 5.76$ volts. This is the point at which the losses would equal the input power; 5.76 volts is also considerably above our anticipated source range. E_b, has been calculated for various values of η_{HRf} and the results have been plotted on Figure 35. A curve has also been plotted on Figure 35 for an optimistic assumption of $P_L = 1.0$ watt/ampere. This is explained in the discussion on page 95 .

C. DETERMINATION OF LIQUID HELIUM REFRIGERANT QUANTITY REQUIRED

Instead of deriving the necessary refrigeration from its output power, the device might be cooled by liquid helium which would be carried as an additional item similar to fuel. This initial estimate will neglect the transducer dissipation and heat lost through the vessel walls which are considerable. The following calculations are based upon the heat leakage through the optimum leads only.

For a device operating at 75% efficiency from a 1.5 volt source and delivering 50 watts the input current is:

$$I_{(input)} = \frac{P_{(output)}}{\eta E_b} = \frac{50 \text{ watts}}{.75 \times 1.5 \text{ volts}} \qquad (135)$$

$$I_{(input)} = 44.4 \text{ amps.}$$

It has been determined in Item B above that the heat leakage per ampere for single stage optimum leads is .084 watt per ampere. For the above device the lead heat leakage Q is:

$$Q = .084 \frac{\text{watt}}{\text{amp}} \times 44.4 \text{ amps} \qquad (136)$$

$$= 3.73 \text{ watts}$$

Since one watt equals .239 gram calorie per second, this heat rate is equivalent to 3.73 x .239 or .893 gram calorie per second. This equals 3600 x .893 or 3218 gram calories per hour. The required weight of helium per hour is:

$$W_{He} = \frac{Q}{K_{He}} \qquad (137)$$

where:

W_{He} = Weight of helium in grams/hr.

Q = Heat rate = 3218 calories/hr.

K_{He} = Latent heat of evaporation of helium

= (5 calories/gram)

Substituting these values in equation (137)

gives:

$$W_{He} = \frac{3218 \text{ calories/hr}}{5 \text{ calories/gram}} = 643 \text{ grams/hr}$$

For an eight hour period this would be 8 x 643 or 5,140 grams. Since the density of liquid helium is 0.122 gram/cc this quantity of refrigerant would occupy a volume of $\frac{5,140 \text{ grams}}{.122 \text{ gram/cc}}$ or 42,100 cubic centimeters.

Expressing the weight in pounds gives

$$W_{He} = 5.140 \text{ Kg} \times \frac{2.2 \text{ lbs}}{\text{Kg}}$$

$$= 11.3 \text{ pounds}$$

Since these calculations consider only one of the losses it is obvious that the actual weight and volume would be much greater than this because the other losses and container must be considered. It can be concluded that the weight and volume of refrigerant required would be excessive for a portable device. Because of this, the approach does not appear practical.

D. CALCULATION OF CRYOTRON TRANSDUCER RATIO REQUIREMENTS:

The transducer efficiency factor η_o can be defined as:

$$\eta_o = \frac{\text{input - transducer loss}}{\text{input}} \tag{138}$$

η_o can be determined for various resistance ratios from the graph of Figure 28.

Equation (138) can be written as:

$$\text{transducer loss} = \text{input}\ (1 - \eta_o) = P_T \tag{139}$$

The efficiency factor for loss dissipated inside the cryotron plus the required compensating refrigeration loss can be defined as follows:

$$\eta_{SRRf} = \frac{\text{input} - (P_T + P_{RFT})}{\text{input}} \tag{140}$$

where:

P_T = transducer loss

P_{RFT} = compensating refrigeration power required (It is assumed that refrigeration power will be obtained from the converted device output at 100% efficiency).

η_{SRRf} = Lumped parameter efficiency factor for above two losses.

It has been found that a minimum of 68.1 watts refrigeration power is required to compensate for each watt of transducer loss. The total minimum refrigeration power required is then:

$$P_{RFT} = 68.1 \text{ watts/watt } (P_T) \tag{141}$$

Substituting equations (139) and (141) in (140)

gives:

$$\eta_{SRRf} = \frac{\text{input} - \left[\text{input}\,(1-\eta_o) + \text{input}\,(1-\eta_o)\,68.1\right]}{\text{input}} \qquad \text{or}$$

$$(\text{input})\,\eta_{SRRf} = \text{input}\left(1 - \left[(1-\eta_o) + (1-\eta_o)\,68.1\right]\right)$$

This reduces to:

$$\eta_{SRRf} = 1 - 69.1\,(1 - \eta_o) \tag{142}$$

It can be noted that the effect of transducer inefficiency is multiplied by 69.1. One of the interesting values is the value of η_o which will result in zero efficiency for η_{SRRf} and hence zero overall efficiency. Thus if $\eta_{SRRf} = 0$, then:

$$\begin{aligned} 0 &= 1 - 69.1\,(1-\eta_o) && \text{or} \\ 69.1 &= 1 + 69.1\,\eta_o && \text{or} \\ 68.1 &= 69.1\,\eta_o && \text{hence} \\ \eta_o &= \frac{68.1}{69.1} \text{ or } 98.6\%. \end{aligned}$$

This indicates that it is necessary to have at least 98.6% transducer efficiency in order to obtain any output whatsoever. Examination of the graph of Figure 28 shows that the resistance ratios for 98.6% efficiency would be off the chart and hence in excess of 10,000 to 1. This ratio would be difficult to obtain when the resistance of optimum header leads and source impedance is considered. A much higher resistance ratio is necessary for any significant efficiency. It can be concluded from this analysis that possible achievement of satisfactory transducer ratios for low voltage conversion does not look very promising. On this basis this approach appears undesirable.

E. CALCULATIONS ON THE SUPERCONDUCTING ELEMENT

Previous calculations have indicated that η_o should equal or exceed .986 to obtain any output whatsoever. Assuming a value of .998 for sample calculations and substituting in equation (139) gives:

$$P_T = \text{Input } (1 - \eta_o) = \text{input } (1 - .998) \tag{139}$$

where:

P_T = Transducer Loss.

For a 50 watt device operating at 75% efficiency the input would be:

$P_{(\text{Input})} = \frac{50}{.75}$ or 66.7 watts.

One might assume for sample calculation purposes that the allowable transducer loss would be:

$$P_T = 66.7\ (1 - .998)$$
$$= .133 \text{ watt.}$$

This loss could also be approximated by:

$$P_T \simeq \frac{(2\ E_b)^2}{R_T} \quad \text{or} \quad R_T \simeq \frac{(2\ E_b)^2}{P_T} \tag{143}$$

where: E_b = Supply voltage.

R_T = Transducer normal resistance

If the supply voltage is .5 volt then:

$R_T \simeq \frac{(2x.5)^2}{.133} \simeq 7.5$ ohms

The required current carrying capacity for a 0.5 volt supply is:

$$I_{(input)} = \frac{P_{(input)}}{E_b} = \frac{66.7 \text{ watts}}{.5 \text{ volt}} \tag{144}$$

$$= 133.3 \text{ amps.}$$

Calculations to determine the physical size of these superconducting transducer conductors follow:

Determine the size of tubular conductor necessary to carry 133.3 amperes. The formula for the magnetic field strength at the surface of such a conductor is:

$$H_c = \frac{2\ Ic}{r} \tag{145}$$

This can be written as:

$$r = \frac{2\ Ic}{H_c} \tag{146}$$

where:

H_c = Critical magnetic field strength in oersteds

Ic = Current in abamperes

r = Radius of the conductor in centimeters

For amperes this becomes:

$$r = \frac{2\ Ic}{10\ H_c}$$

The critical field strength is normally expressed in gauss in place of oersteds. The permeability of free space is 1 gauss/oersted. A value of 200 oersted or 200 gauss may be reasonable.

Thus

$$r = \frac{2 \times 133.3}{10 \times 200}$$

$$= .1333 \text{ centimeter}$$

Thus the conductor should have a diameter of .267 cm. This should be a tubular thin film conductor in order to maintain high resistance in the non-superconducting state. The minimum practical film thickness is 5×10^{-6} cm. The required length of the conductor can be determined from:

$$\ell = \frac{R_T \pi D t}{\rho}$$

where:

ℓ = Conductor length

R_T = Normal resistance = 7.5 ohms

ρ = Resistivity = 10^{-7} ohm -cm

D = Conductor diameter = .267 cm

t = Film thickness = 5×10^{-6} cm.

Substituting these values gives:

$$\ell = \frac{7.5 \pi (.267) \; 5 \times 10^{-6}}{10^{-7}} \quad \text{or}$$

$$\ell = 31.5 \text{ centimeters}$$

This dimension is reasonable although the conductor may have to be coiled or folded back upon itself. The conductor might consist of a thin film deposited on a round insulating rod or tube. It can be noted that the allowable conductor diameter will change directly with the required current capacity. The current required is inversely proportional to the voltage. The resistance required is directly proportional to the square of the voltage, however. From this it can be noted that the conductor length will be directly proportional to the input voltage. Thus for a 1/2 volt input the conductor length is 31.5 cm. Construction of a device to handle the higher input voltages would become more complex due to the increased conductor length required. These calculations indicate that fabrication of a low input voltage device would require a conductor of relatively reasonable dimensions whereas the higher input voltage device would be more difficult because of the long conductor required. Our other calculations show that the low input voltage device is not feasible because of the high header lead heat flow and refrigeration losses.

The resistance ratio of the above conductor can be estimated if it is assumed that one half the .084 watt per ampere of heat transfer in the optimum header lead is caused by I^2R loss. On this basis the I^2R loss is .042 watt per ampere. For 133.3 amps this is 133.3 amps x .042 watt/amp or 5.6 watts $= P_L$. Also:

$$P_L = I^2 R_L; \text{ or} \qquad (148)$$

$$R_L = \frac{P_L}{I^2}$$

where: R_L = resistance of input lead pair

Substituting values in (148) gives

$$R_L = \frac{5.6}{(133.3)^2}$$

$$= 3.14 \times 10^{-4} \text{ ohms.}$$

Since the resistance of the element is zero when superconducting and 7.5 ohms when normal the resistance ratio is:

$$\frac{R1}{R2} = \frac{R_L + R_{(\text{element normal})}}{R_1 + R_{(\text{element superconducting})}} \qquad (149)$$

$$\simeq \frac{R_{(\text{element normal})}}{R_L}$$

$$\simeq \frac{7.5 \text{ ohms}}{3.14 \times 10^{-4} \text{ ohms}}$$

$$\frac{R1}{R2} \simeq 23,800$$

This is a respectable ratio which tends to approach the desired value. These calculations indicate that the conductor dimension may be feasible for low input voltages. Other conditions mentioned above, however, tend to rule out feasibility for low voltages.

F. CRYOTRON CONVERSION

We explicitly consider here the application of cryotron switching to the push-pull converter shown in Figures 27, 28. A schematic illustration is given in Figure 34. Several film cryotrons are connected in series. Use of film cryotrons instead of wirewound cryotrons anticipates the need for higher resistivity in this application. The means of alternate switching between top and bottom arrays is not described here. A method for accomplishing this is inherent in the cryotron multi-vibrator described by Buck.*

1. Required Gate Geometry

R_2 is the normal resistance of the cryotron gate and R_1 the internal resistance of the source plus lead resistance. We will, in fact, assume that the source resistance is negligible. In computing the lead resistance it must be assumed that the source is outside the cryostat, and that the leads extend from room temperature to liquid helium temperature ($4.2^{o}K$). As shown in previous calculations the desired value of R_2 is 7.5 ohms.

The materials available for the gate are mainly tin, tantalum, and lead (Pb). In general the "soft" superconductors are more useful for this purpose because of low critical fields. The low temperature normal resistivity of impure specimens of these metals is typically 10^{-7}. The gate resistance is then given approximately by:

$$R_2 = \rho \frac{w}{Wd} \tag{150}$$

where:
- R_2 = Normal gate resistance = 7.5 ohms
- W = Gate width (assume 2 cm.),
- d = Gate thickness (5×10^{-6} cm, (close to the feasible minimum)).
- w = Control width
- ρ = Normal resistivity (assume 10^{-7} ohm cm)

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* D. A. Buck, "Cryotron - A Superconductive Computer Component," I.R.E. Proc. 44:482-93, April 1956.

We find,

$$R_2 = \rho \frac{w}{Wd} = \frac{10^{-7} w}{2x5x10^{-6}}$$

$$R_2 = w\, 10^{-2} \text{ ohms/cm}$$

Since $R_2 = 7.5$ ohms

$$w = \frac{7.5 \text{ ohms}}{10^{-2} \text{ ohms/cm}} = 750 \text{ cm.}$$

Thus the required cryotron gate length is estimated to be at least 7.5 meters. This would, of course, be divided into separate cryotrons connected in series as shown in Figure 34. The number of such cryotrons would appear to be large, as shown in the following section.

2. Power Consumption

The control current required for operation of a shielded crossed-film cryotron has been shown by Newhouse* to be:

$$I_c \text{ (amps.)} = 10\, H_F\, w/4\pi \tag{151}$$

where H_F is the critical field of the gate film in oersteds and w is in cm. If H_F is taken as 10 oersteds and I_C = 10 amps (an appreciably greater current into the cryostat would result in a rather large consumption of liquid helium) the resulting gate length per cryotron is about 1.26 cm. The required number of cryotrons would then be 750/1.26 or 595. At the present time this appears very complex.

The energy consumed in switching a cryotron is equal to the magnetic energy stored in turning on the current and is thus given by $1/2\, LI_c^2$, where L is the self-inductance of the shielded control film. The latter is given by

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* V. L. Newhouse, "Superconductive Circuits for Computing Machines", Electro-Technology, 67:78-89, April 1961.

$\left[4\pi tW/w\right] \times 10^{-9}$ henries where t is the separation (cm) between control film and shield plane, W and w having the usual meaning. Finally, if ϑ is the switching frequency, the cryotron power is given by:

$$P_{watts} = \vartheta W t H_f^2 \times 10^{-7} / (8\pi) \qquad (152)$$

For $\vartheta = 1000$ sec^{-1} and t $\simeq 10^{-4}$ cm, P $\simeq 3.98 \times 10^{-8}$ watts. The total power consumption under these conditions would then be 595x3.98 x 10^{-8} or about 24 microwatts for the converter.

Appendix D Prepared by O. Lutes

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APPENDIX E

BIBLIOGRAPHY

A bibliography has been prepared as part of the literature search. Most of the literature listed has been ordered, but not all of it has been received and examined.

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LOW INPUT VOLTAGE CONVERSION - by J. T. Lingle, C. Motchenbacher, D. Long, O. Tufte, D. Benz, O. Lutes, J. Garfunkel

Quarterly Report No. 1. July 1 to September 30, 1962. 219pp., 49 illus., 9 tables.

Contract DA-36-039-SC-90808. Department of Army Project No. 3A99-09-001. Unclassified Report

A literature search was made to gather information on all known methods, techniques, and circuitry for conversion of low d-c input voltages from thermoelectric, thermionic, solar cells, fuel cells, and other single cell sources to a higher, more usable voltage. Analytical feasibility studies were conducted on transistor, tunnel diode, electromechanical, Hall effect, magnetoresistive, superconductive, and photoresistive methods of low input voltage power conversion. The parameters required for transistor, tunnel diode, and other transducers were calculated.

Transistors selected from present production units might be satisfactory for construction of converters operating from a 1.5 volt source. For operation from source voltages less than 1 volt, special transistors would have to be developed. The use of a transistor converter having both voltage and current feedback drive appears desirable. The use of four or more transistors appears desirable in order to obtain the desired efficiency and power output.

(over)

Unclassified

1. Low Input Voltage Converters - Methods - Feasibility.
2. Semiconductors
3. Circuits
4. Magnetic Properties
5. Electrical Properties
6. Photoelectric Materials
7. Resistance (Electrical)
8. Superconductivity

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2. Semiconductors
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4. Magnetic Properties
5. Electrical Properties
6. Photoelectric Materials
7. Resistance (Electrical)
8. Superconductivity

I. Title: Power Supplies
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